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Journal of Communications and Electronic Engineering

(Including the WAVES AND ELECTRONS Section)

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Volume 37 Number 8

Sylvania Electric Products Inc.

TOMORROW'S TUBES AND CIRCUITS

formance characteristics of the traveling-wave tube, and its asso-uit "plumbing," are under engineering study.

PROCEEDINGS OF THE I.R.E.

Path-Length Microwave Lenses

Mercury Delay Line Memory with Megacycle Pulse Rate

Magnetic Amplifiers with Feedback

Part II—Electrical Network Analyzers for Electromagnetic Field Problems

Optimum Transient Response Amplifiers

Microwave Switching Tube Admittance

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Technical Memoranda Cover Sheet for Information Exchange

Radioactivity Measuring Instruments and Standards

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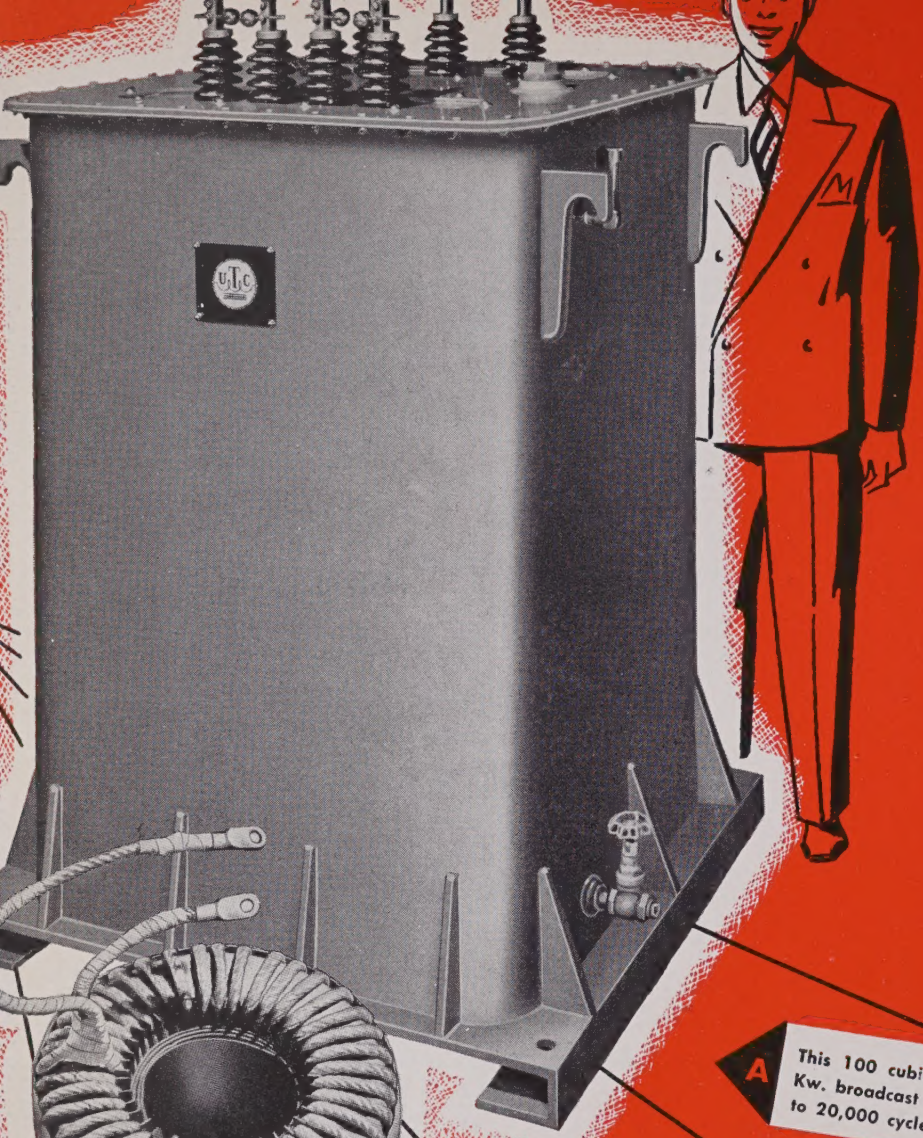

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William L. Everitt

BOARD OF DIRECTORS, 1949–1951

William Litell Everitt was born on April 14, 1900, in Baltimore, Maryland. During World War I he served in the U. S. Marine Corps from 1918 to 1919. At the war's conclusion, he matriculated at Cornell University, where he taught electrical engineering from 1920 until he received the E.E. degree in 1922.

In that year he joined the North Electric Manufacturing Company of Galion, Ohio, as engineer in charge of the design and development of their relay automatic public switchboard exchanges. He left in 1924 to teach electrical engineering at the University of Michigan, transferring to Ohio State University in 1926—the year in which he received the M.A. from Michigan—to take charge of their communications engineering staff in the capacity of assistant professor. Meanwhile, during the summers from 1925 to 1930, he served with the department of development and research of the American Telephone and Telegraph Company.

Joining The Institute of Radio Engineers as an Associate Member in 1925, he became a Member in 1929, and also was elevated to the rank of associate professor at Ohio State in that year. In 1933 he received the Ph.D. degree and was promoted to a full professorship. At Ohio State he originated and directed the annual Broadcast Engineering Conference, in which the IRE partic pated. He became a Fellow of the IRE in 1938.

In 1940 Dr. Everitt was appointed a member of th Communications Section of the National Defense R search Committee. Two years later he took a lea of absence from the University to serve as director operational research with the U. S. Army Signal Corp He was appointed head of the University of Illinoi electrical engineering department *in absentia* in 194 and at the war's conclusion he took up his duties ther Effective September 1, 1949, he will become Dean of th University of Illinois' engineering college.

The author and editor of numerous texts and articl on electrical engineering, Dr. Everitt has been a co sultant for various broadcast stations and radio manu facturing companies. He was President of the IRE i 1945 and has been a member of a number of committee currently he is on the Circuits Committee and th Board of Editors. Dr. Everitt is also a Fellow of th AIEE, a member of the National Council of Tau Bet Pi, and a member of Sigma Xi, Eta Kappa Nu, th Acoustical Society of America, the American Societ for Engineering Education, and the American Associ tion for the Advancement of Science.

The Valuable By-Product

V. K. ZWORYKIN

Scientists, in their systematic study of phenomena and their development of the covering theories, pursue a lengthy and sometimes tortuous path. Shall they always head unswervingly toward the distant goal or should they occasionally pursue, or at least point out, interesting side paths?

These matters are considered in the following guest editorial by an eminent research worker, the correctness and fruitfulness of whose methods have been amply proven. He is a vice-president and technical consultant of the RCA Laboratories, and a Fellow of the IRE and member of the Board of Editors.—*The Editor.*

"Select your goal carefully, and then, looking neither to left nor right, pursue it."

That is good advice. Yet, like most good advice, it can be overstressed. Much that is of great value has come from the exploration of bypaths in science and engineering. Certainly, few will dispute Röntgen's wisdom in ferreting out the cause of the lighting up of a fluorescent screen near his enclosed gas discharge tube, and studying the phenomenon with such thoroughness that he was able not only to announce to the world the existence of X rays, but also to give an extraordinarily complete description of their characteristics. Perhaps the ideal compromise between persistence in a chosen line of investigation and the exploration of an incidental find was practiced by Hertz in the discovery of the photoelectric effect. Having noted unexplained nonuniformities in the length of induced sparks, he performed a series of critical experiments with available equipment, which demonstrated that these arose from the action of ultraviolet light on the negative spark electrode. After publishing a careful description of these experiments for the guidance of later investigators, Hertz returned to his original objective—the demonstration of electric waves as predicted by Maxwell's theory—and thus helped to lay the basic groundwork for the radio industry.

There are many examples of the valuable by-product in our own experience in the development of television. Perhaps the first that comes to mind is the multiplier phototube which now plays such a vital role as detector in nuclear and cosmic-ray work, and has found innumerable other applications in industry and research. Our immediate interest in secondary-emission multiplication stemmed from the need of increasing the signal strength of television pickup tubes. Again, the electron microscope, which has extended the range of our vision by two orders of magnitude, naturally fitted into the orbit of our interests as the result of our intensive occupation with the electron-optical problems of television; the circuit experience gained in television development helped to create an instrument of compact dimension with a resolving power approaching the theoretical limit. The infrared-sensitive image tube, which came to play an important part as reconnaissance instrument during the war, was another early by-product of pick-up tube research. And now, a variety of storage tubes, the offspring of television research in tube construction, operating principles, and circuitry, promise to make material contributions to air safety through systems such as Teleran, to facilitate the study of high-speed phenomena, and to bring within reach the solution of mathematical problems of baffling complexity by modern electronic computing machines.

The list is not complete, nor is there any sign of a halt in the flow of new ideas and new devices that spring from television development. There can be little question that the same applies to any number of other lines of endeavor. It is with this in mind that I would change the advice given above to the following:

"Select your goal carefully, and then, looking both left and right, proceed on your way, marking out promising side paths as they meet your eyes. Thus, you may reach your goal sooner and help others, in turn, to reach their goals."

Path-Length Microwave Lenses*

WINSTON E. KOCK†, SENIOR MEMBER, IRE

Summary—**Lens antennas for microwave applications are described which produce a focusing effect by physically increasing the path lengths, compared to free space, of radio waves passing through the lens. This is accomplished by means of baffle plates which extend parallel to the magnetic vector, and which are either tilted or bent into serpentine shape so as to force the waves to travel the longer-inclined or serpentine path. The three-dimensional contour of the plate array is shaped to correspond to a convex lens. The advantages over previous metallic lenses are: broader band performance, greater simplicity, and less severe tolerances.**

Introduction

LENSES ARE useful microwave antennas because of their tolerance advantages. An amount of twist or warp of the completed antenna which would seriously degrade the gain and directional properties of a parabolic dish reflector antenna will generally have negligible effect upon the performance of an equivalent lens.

Two types of metallic microwave lenses developed by the Bell Telephone Laboratories have previously been described. The first comprised rows of conducting plates which acted as waveguides and achieved a focusing effect by virtue of the higher phase velocity of electromagnetic waves passing between the plates.[1]

The second type was a scaled-up version of the lattice structure of a true dielectric, whereby small conductin elements replaced the molecules of the dielectric, an the polarization which these elements engendered dupl cated the polarization of the true dielectric.[2] This type lens exhibited very broad frequency characteristics, that it was effective over a much larger wavelength ban than the previous waveguide lens. Because, howeve the elements were small compared to the wavelengt large size lenses of this type contained large numbers the conducting elements, and the construction becam somewhat tedious. For example, the partially assemble 10-foot delay lens shown in Fig. 1 (strip type structur requires several thousand strips to fill the 10 foot ape ture. The lenses to be described in this paper retain th broad-band features of the delay type, but permit simpler construction.

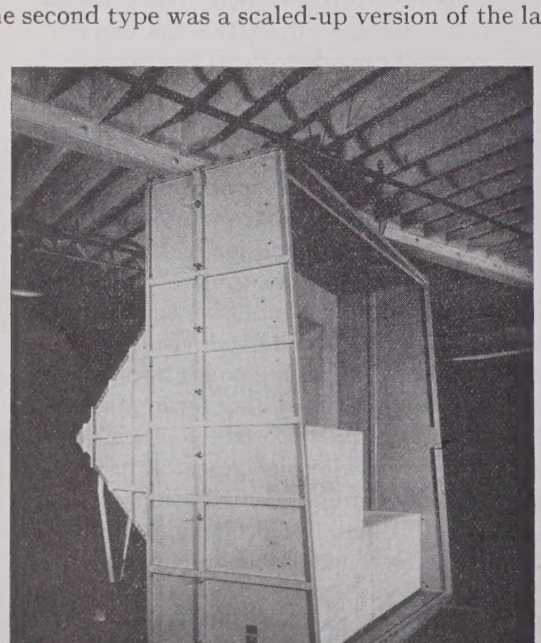

Fig. 1—A partially assembled 10-foot artificial dielectric shielded-lens antenna for microwave relay use.

Fundamental Principles

The principle of operation of the path-length lens can be described in connection with Fig. 2. If parall

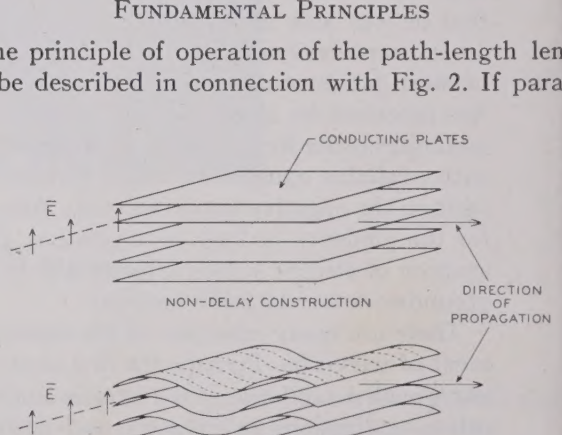

Fig. 2—Use of conducting plates or grids (shown dotted) to effect a wave delay.

conducting plates are presented to electromagnet waves polarized perpendicularly to the plates, little fect will be produced on the progress of the waves, pr viding that the plates are flat and aligned along the rection of propagation, as shown in the top of the figu If, however, the plates are bent into serpentine shape, shown in the bottom of the figure, the sinuous path l, i side the plates, will be longer than that outside l_0, a delay will be produced. In order to avoid the generati of a second-order mode, the vertical spacing of the plat must be less than one-half wavelength, but for all wav lengths longer than this, a constant delay and thus constant refractive index is obtained. If, instead of t serpentine construction, the plates are tilted so th they form an angle with the direction of propagation delay will also be produced, since the waves will be forc

* Decimal classification: R326.8×R310. Original manuscript received by the Institute, January 11, 1949.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] W. E. Kock, "Metal-lens antennas," PROC. I.R.E., vol. 34, pp. 828–837; November, 1946.

[2] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, 27, pp. 58–83; January, 1948.

traverse the longer inclined path. Instead of solid ates, a grid structure can be used if desired, as shown y the dotted lines in the bottom of Fig. 2. The wires of ie grid must be less than one-half wavelength apart in rder to simulate the effect of a solid sheet.

Design Considerations

The design of microwave lenses using the path-length rinciple follows, in general, the design of any lens in hich the material comprising the lens has an index of fraction greater than unity. For example, in Fig. 3, the ntour *A CGHD* defines the cross section of a slant plate ns looking end on at the plates, and the curved portion *CG* is obtained by making the path lengths of all rays ual from the focus to the plane *BEI*. The added delay increased path length of ray 2 is equal to the added ngth of the line *CD*, compared to its free-space path he thickness of the lens from point *C* to the front sur- ce). The effective index of refraction is thus equal to /cosine θ.

When the lens has a flat front surface, the curved side, hich is towards the feed, turns out to be a hyperboloid revolution, as in the usual case of dielectric lens. The uation of the generating hyperbola of this surface is[3]

$$(n^2 - 1)x^2 + 2fx(n - 1) - y^2 = 0, \qquad (1)$$

here n is the effective index of refraction ($1/\cos\theta$ for ie slant plate and l/l_0 for the serpentine lens), f is the cal length, and the origin of the co-ordinate system is the point C in Fig. 3.

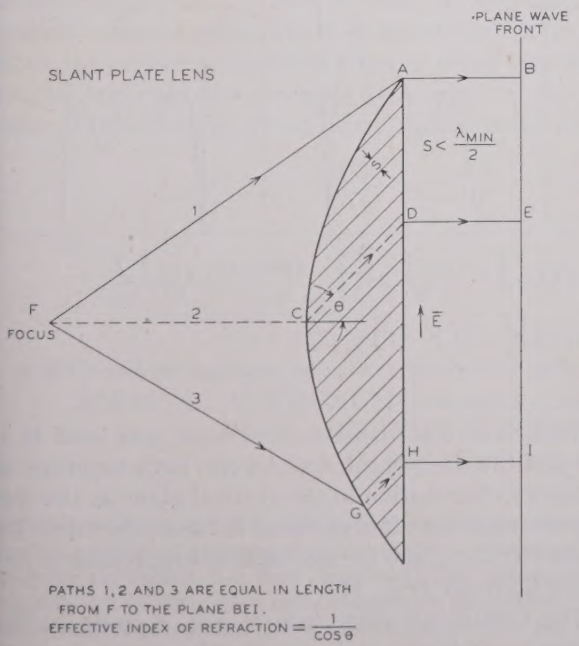

Fig. 3—Determinaton of lens contour for path-length lenses of the slant-plate variety.

For the lens reversed, so that rays starting from the cal point strike the flat surface, the contour of the outer curved surface must assume an asymmetrical shape in order that a plane wave emerge from it.

Construction Methods

Two types of lightweight grid structures are shown in Fig. 4. In both cases, the vertical spacing b and the horizontal spacing a of the grid wires must be less than one-half wavelength for proper operation, as explained earlier. The serpentine or slanted wires are affixed to polystyrene foam slabs, and the contour of the wire grid outline on each slab is a section parallel to the axis of the hyperboloid of revolution generated by (1). Polystyrene foam (Styrofoam, Dow), is light in weight and transparent to microwaves.

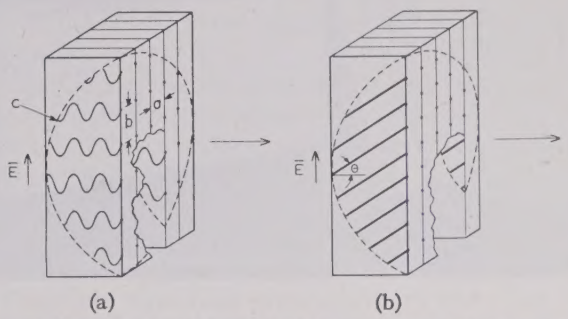

Fig. 4—Wire grid structures for path-length lenses. (a) Serpentine grid construction; (b) slant grid construction.

A photograph of a serpentine lens of the type sketched in Fig. 4(a) is shown in Fig. 5. This lens is 12 inches square and was designed for 3-cm microwaves. A two-foot square slanted wire lens of the type sketched in Fig. 4(b) is shown in Fig. 6. It was designed for use at 7-cm wavelengths. One of the foam slabs with grid wires attached is shown separately, and the remaining slabs have not yet been put into the frame.

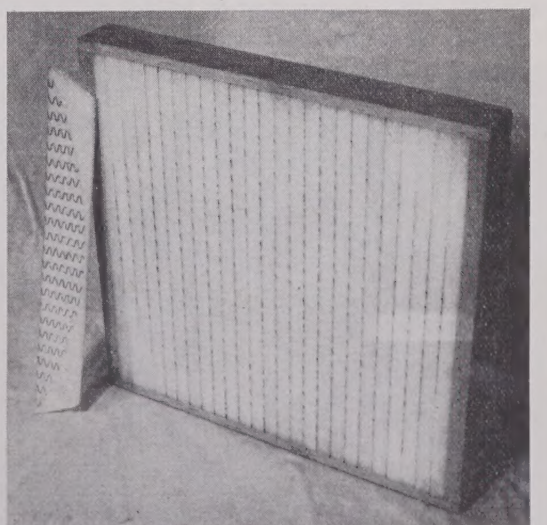

Fig. 5—One-foot diameter serpentine wire lens for 3-cm wavelengths.

Fig. 7 shows the magnetic plane pattern of this lens when illuminated with a small feed horn. The feed horn

[3] See Fig. 17 in footnote reference 2.

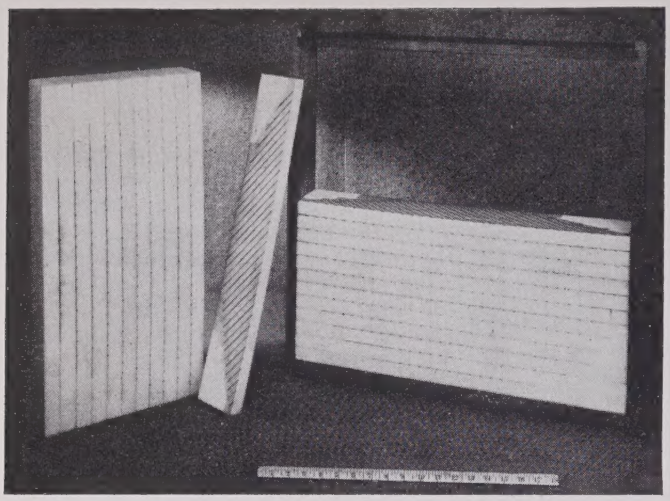

Fig. 6—A partially assembled slant wire lens for 7-cm operation.

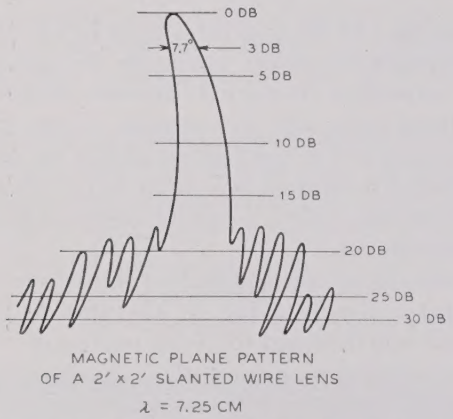

Fig. 7—Radiation pattern of the lens of Fig. 6.

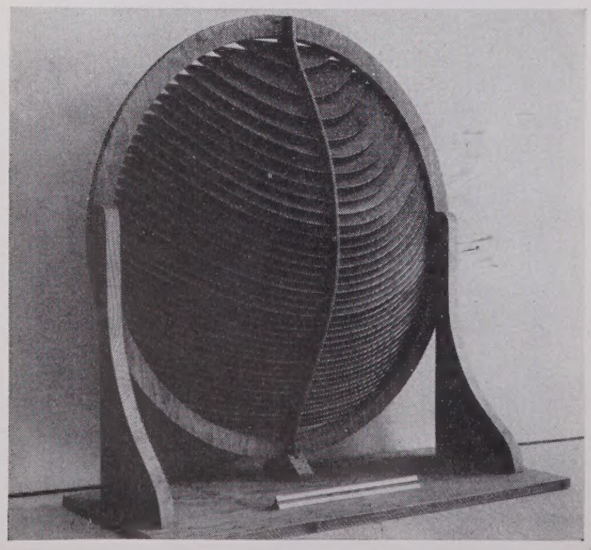

Fig. 8—Continuous sheet slant plate lens 30 inches in diameter.

had insufficient directivity to suppress the minor lobes as well as might be desired, but the symmetry and beam width of the pattern indicate satisfactory operation the lens.

For slant plate lenses employing solid plates instea of grids, the profiles of the plates again are sections the design hyperboloid, but they are now formed b planes making an angle θ with the axis of revolutio This is shown in Fig. 8, where a 30-inch diameter ope structure solid plate lens for use at microwavelengt from 2.5 cm upward is portrayed. In this lens, aluminu plates are supported at the edges and spaced by an i sulating member along the central plane. It has a foc length (f in equation (1)) of 30 inches, and an angle tilt of the plates of 48.2 degrees, corresponding to a index of refraction (n in equation (1)) of 1.5. Because its open structure, air can pass readily between th plates.

A magnetic plane pattern of this lens is shown in Fi 9, and it is seen that the minor lobes are fairly well su

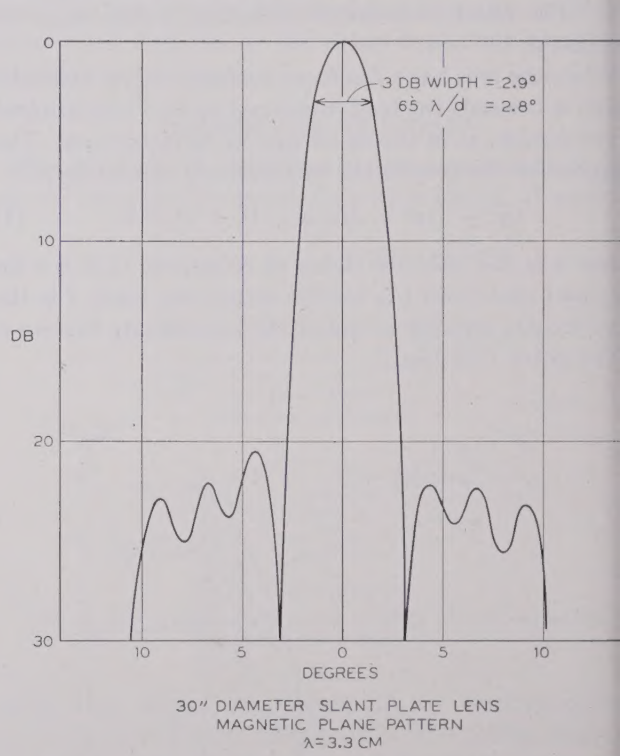

Fig. 9—Horizontal radiation pattern of the lens of Fig. 8.

pressed. A 3-inch diameter feed horn was used in th test and the wavelength was 3.3 cm. Because of the shi in energy distribution in the vertical plane as the wav pass through the tilted plates of the lens, the minor lob in the electric plane are not as well suppressed as in th magnetic plane (Fig. 10).

The continuous plate lens of Fig. 8 could be ma lighter in weight by replacing the aluminum plates wi copper foil sheets of proper contour, and then employi foam slabs as spacers and supports between the plate Currents flow along lines lying in vertical planes, so th the lens could be sectioned vertically for assembly pu poses in the case of large structures.

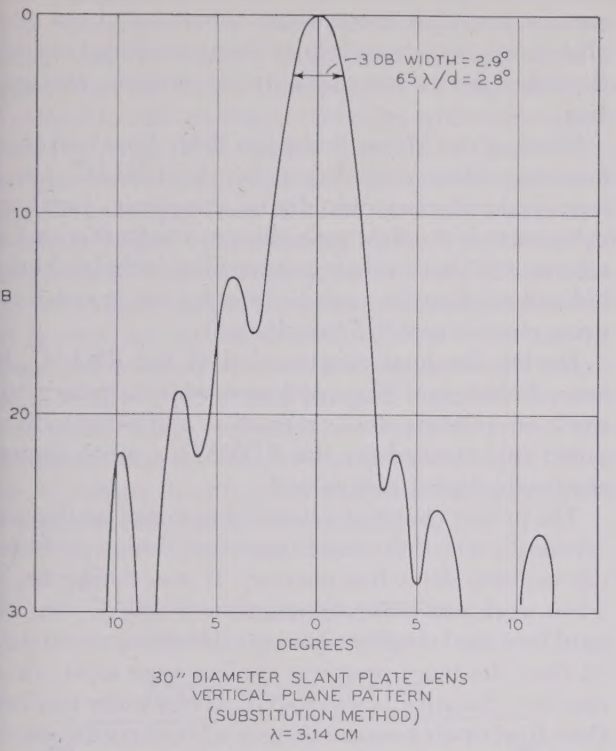

Fig. 10—Vertical radiation pattern of the lens of Fig. 8.

Comparisons with Earlier Types

Because the path-length lenses function identically for ll wavelengths longer than the second mode wave-ength, the index of refraction remains constant up to his wavelength limit. There is thus no dispersion or hange of index of refraction with wavelength, as in the ase of the earlier conducting element delay lens when perated too near the resonant frequency of the ele-ients. If enclosed in a full horn shield, such a lens should, over its entire operating range, possess an effective area which is independent of frequency. It would thus constitute an extremely broad-band antenna.

Since the index of refraction of the slanted plate lens depends only upon the tilt of the plates and not upon the plate spacing as in the first waveguide lens, constructional tolerances of this lens are even less severe than in the earlier lenses. Thus the number of plates could be doubled, if desired, without altering its performance in the original operating band. Also, flatness of the plates is not important, so long as the second mode spacing is not exceeded. The simplicity of design is brought out by observing that only 60 flat sheets cut to proper profile are required to duplicate the effectiveness of the several thousand elements of the 10-foot lens of Fig. 1.

There are two disadvantages which the slant plate lens possesses. The first is that the lens plates must be held at the proper design tilt with respect to the feed horn (since the index of refraction depends upon plate tilt). This prohibits a lens tilt in the vertical plane unless so designed, and limits the scanning ability of a moving feed in the vertical plane. The usual lens scanning capabilities are retained in the horizontal plane, however. The second is that the energy distribution in the electric plane is unsymmetrical, and results in a poorer minor lobe suppression in this plane. This latter disadvantage may not be serious in applications such as microwave repeater work, where lobes in the vertical plane are not as objectionable as lobes in the horizontal plane.

Acknowledgment

The writer wishes to acknowledge the assistance of W. E. Legg in the construction and testing of the lenses discussed in this paper.

Mercury Delay Line Memory Using a Pulse Rate of Several Megacycles*

ISAAC L. AUERBACH†, member, ire, J. PRESPER ECKERT, Jr.‡, associate, ire, ROBERT F. SHAW‡, senior member, ire, and C. BRADFORD SHEPPARD‡, member, ire

Summary—**A mercury delay line memory system for electronic omputers, capable of operating at pulse repetition rates of several egacycles per second, has been developed. The high repetition te results in a saving in space and a reduction in access time.**

Numerous improvements in techniques have made the high epetition rate possible. The use of the pulse envelope system of epresenting data has effectively doubled the possible pulse rate; the use of crystal gating circuits has made possible the control of signals at high pulse rates; and a multichannel memory using a single pool of mercury has simplified the mechanical construction, reduced the size, and made temperature control much easier.

The memory system described makes possible a significant increase in the over-all speed of an electronic computer.

* Decimal classification: 621.375.2×R117.19. Original manuscript received by the Institute, December 15, 1948; revised manuscript received, March 4, 1949.

† Formerly, Eckert-Mauchly Computer Corp.; now, Burroughs dding Machine Co., Philadelphia, Pa.

‡ Eckert-Mauchly Computer Corp., Philadelphia, Pa.

I. Introduction

THE MAJOR design problem in any high-speed electronic computer of advanced design is that of a memory, or storage device. Such a device must be capable of storing several hundred or more numbers

of, say, twelve decimal digits each. Furthermore, any one of these numbers must be available for use in the computing circuits within a time comparable to the time required for an elementary arithmetic operation. Although considerable research is being done on electrostatic memories, in which information is stored in the form of charges on a dielectric medium, the delay-line type of memory is more highly developed at the present time, and is being used in several computers. In a delay-line memory, information is stored in the form of groups of electrical or acoustical impulses or signals circulating in an electric delay line or medium suitable for transmission of acoustic waves.

A device capable of storing n pulses may also be considered as storing n binary digits. The presence or the absence of a pulse in each of the n successive positions provides the coding necessary to represent binary numbers. To represent a decimal digit, a minimum of four successive coded pulse positions is required.

Acoustic propagation has a lower velocity than electromagnetic propagation, so a proportionately greater amount of information can be stored in the acoustic type delay line than in the electrical delay line for a given delay bandwidth and attenuation. Where large storage capacity is required, the acoustical line is preferred. Because of the low velocity of propagation and the good impedance match between it and quartz, mercury has been selected as the acoustic delay medium.

II. History

One of the first applications of acoustic delay lines was in the Scophony television system,[1] where video signals were converted into acoustical wave patterns in a transparent medium. These signals were read out optically by making use of the Debye-Sears effect[2] which translates the strain pattern into a pattern of varying light intensity.

This type of acoustic line was a nonregenerative or "delay storage" device. Further work on acoustic lines was done by Shockley at Bell Telephone Laboratories, who also used them as simple time delay devices. In 1943, Eckert, then at the Moore School of Electrical Engineering at the University of Pennsylvania, carried on further development work under a contract from the Radiation Laboratory of the Massachusetts Institute of Technology. Eckert was the first to use mercury as the acoustic medium and the first to use nonreflective backings on the quartz crystal transducers. By this technique it was possible to eliminate multiple storage due to reflections, and it was no longer necessary to use amplitude discrimination at the receiving end, as had been necessary in Shockley's lines. It also became possible to store patterns of pulses with relatively close spacing, increasing the amount of information per unit length, due to the greater bandwidth which the nonreflective terminations provided. Once these things had been accomplished, regeneration circuits were introduced to convert the delay device into a long-time dynamic storage device.

Work at the Moore School on delay lines had to stop however, when work began on the ENIAC, the first large-scale all-electronic digital computer. Further development in the field took place at the Radiation Laboratories.[3–5] These subsequent studies included both liquid and solid media acoustic lines for use in range measuring circuits and MTI equipment.

During the final construction of the ENIAC, however, Eckert and Sheppard were able to return to the work on acoustic delay memories and eventually proposed this method for the EDVAC, a much improved electronic digital computer.[6]

The proper choice of a backing material for the quartz crystals is one of the most important design problems in the mercury delay line memory. It was during the time when work was being done on the EDVAC that Sheppard first used steatite as a crystal backing material. Until then the more practical choices were steel, air, and mercury. Steatite, it was found, gave a wider bandwidth than steel or air because there is a lower coefficient of reflection between it and quartz. The mercury-backed crystal had a higher capacitance than that with air backing, and further, this mounting subjected the crystals to breakage due to hydrostatic pressure of the mercury, and due to the method of mounting the thin crystals.

III. Theory

Delay line circulating memories have one common characteristic: an inevitable distortion of the circulating pulses. The pulses will be distorted in shape, and a timing shift will occur after each circulation. The shape distortion is due to attenuation, dispersion, phase distortion and other less important practical considerations. If the shape distortion were not corrected, the pulse group would lose their identity after a very few circulations. The timing shift arises through small errors produced in the temperature control system, changes in the length of the mercury column due to expansion and contraction and the inaccuracy of construction which may give the different channels slightly different lengths. If the timing shift were not held within limits, it would be difficult to select a specific group of pulses from the several circulating in a single channel. Additional difficulties would arise in the arithmetic circuits, since timing synchronism is essential for proper operation. If the operation of the line is to be successful, both distortion and timing shift must be corrected.

[1] V. K. Zworykin and G. A. Morton, "Television," John Wiley and Sons, Inc., New York, N. Y., p. 254; 1940.
[2] See p. 251 of footnote reference 1.
[3] "A Theory of the Supersonic Delay Line," Rad. Lab. Report No. 733.
[4] "Multiple Reflection Delay Tank," Rad. Lab. Report No. 7
[5] "On the Theory and Performance on Liquid Delay Lines," Rad. Lab. Report No. 792.
[6] "Progress Report on the Edvac," University of Pennsylvania, Moore School of Electrical Engineering, June 30, 1946.

The actual processes for correcting these variations may be thought of in two different ways. The terms "reshaping" and "retiming" imply that each pulse, as it emerges from the line, is sharpened to its original shape and shifted in time so that it coincides with some standard timing pulse. It is equally possible to think of the distortion correction process as being one in which each pulse, as it emerges from the line, is used to operate a gate circuit which, upon opening, allows a new standard timing pulse of the proper shape to enter the acoustic delay line. In this sense the old pulse containing some small error in time, and having been broadened by its passage through the memory channel and its associated amplifiers, is finally used to hold the gate open for the new timing pulse. Each recirculation therefore uses a new pulse gated into the line by the previous pulse. If no pulse is present in any given position, the gate circuit will suppress the standard timing signal.

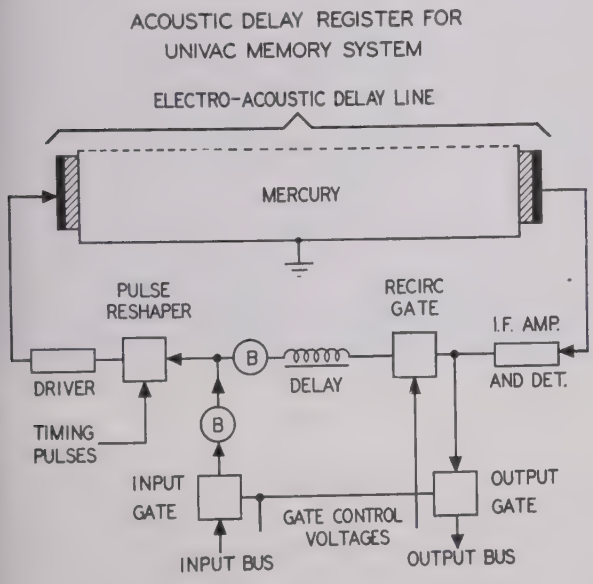

Fig. 1

Fig. 1 shows a block diagram of a typical memory channel. Pulses from the input bus enter the recirculation circuit through the input gate if an input gating signal is supplied. These pulses then gate timing pulses through the pulse reshaper or clock gate. The output of the clock gate feeds the driver, which is simply a power amplifier. Upon leaving the amplifier, the electrical pulse is applied to a quartz crystal transducer at one end of the mercury column, causing the transducer to produce an acoustic pulse in the mercury. The acoustic pulse then travels through the mercury with the velocity of sound in that medium. Because of the bandwidth limitation of the mercury tank, the acoustical pulse, after conversion into an electrical signal, has become a wave packet. After being amplified in the band-pass amplifier, the signal is rectified by the detector. The band-pass amplifier has a center frequency comparable to the crystal frequency and a bandwidth of several megacycles. The bandwidth limitation of the amplifier further broadens the wave packet. The output of the detector is a signal whose shape is that of the envelope of the wave packet. This signal, after passing through the recirculation gate, is used to gate a new timing pulse through the clock gate. The process is repeated indefinitely, unless the line is cleared by interrupting the recirculation path or a power failure occurs.

The two most important considerations of such a memory device as is here described, are the amount of information which it can store and the average length of time needed to remove or insert a given group of pulses. The waiting or latency time, often called the access time, directly affects the speed of operation of the computer. On the average, the access time will be one-half the recirculation time. It is as probable that a group of pulses will be desired just after they have entered the memory channel as it is that they will be desired just as they are about to emerge. In view of the access time consideration, several short channels are to be preferred over a few long ones.

The capacity of a delay line memory is determined by the length of the column and the repetition rate of the pulses circulating in the memory. Increasing either will increase the capacity, but it has already been pointed out that column length should be kept short in the interests of access time. Therefore, larger memory capacity should preferably be achieved through increased repetition rate. Considered in another way, a higher repetition rate will, for a given number of pulses per channel, reduce the access time because this number of pulses can be put in a shorter length of column. Any shortening of the access time makes it reasonable to increase the speed of performing arithmetic operations within the computer.

An important factor in determining the maximum repetition rate is the dispersion or phase distortion resulting largely from the effect of the walls of the tubes containing the mercury column. Such effects are greatly minimized by making the diameter of the column large compared to the crystal diameter and compared to the wavelength. Using crystal frequencies of about 15 Mc, this is not difficult, but still would result in a rather bulky memory assembly if the number of columns were large. It has been found, however, that a common pool of mercury can be used for a number of memory channels. The acoustical waves are so directional that each crystal affects only the one directly opposite it; and cross talk between channels can easily be held to two or three per cent with columns long enough to store over a thousand pulses per channel, particularly, if stainless steel tubes or "liners" such as those shown in Fig. 4 are used to isolate some channels from their neighbors.

The attenuation in the mercury has been found to be

$$\alpha = 0.0013f^2 + 0.9/fd^2,$$

where

α is the attenuation in db per hundred microseconds of delay

f is the frequency in megacycles

d is the column diameter in inches.

Having made the second term negligible by using a large column diameter, one is still faced with the problem of frequency discrimination. The bandwidth required depends only on the repetition rate of the pulses; by increasing the center frequency of the pass band, the variation of α over a given bandwidth can be decreased.

If effects of dispersion and attenuation are minimized, as described, the chief limitation on bandwidth is the electromechanical coupling characteristic of the crystals. The response is proportional to $\sin^2 \phi/\phi$, where ϕ is the ratio of impressed frequency to crystal frequency, and the distance between the half-power points $\Delta f = 0.66 f_c$, where f_c is the crystal frequency, assuming proper acoustic termination.[7] To find the required bandwidth for transmission of the wave packets, it will be assumed that the crystal frequency is made equal to about three times the maximum pulse repetition rate; it is then reasonable to use the Fourier transform of the envelope of the packet to obtain the frequency spectrum. Assuming an envelope having the shape of the probability curve, given by

$$\frac{\epsilon(t)}{e_{\max}} = \epsilon^{-2.8(t/t_0)^2},$$

the envelope of the corresponding frequency spectrum is given by[8]

$$\frac{e(f)}{e_{\max}} = \frac{1.1 t_0}{T} \epsilon^{-3.6(f-f_p)^2/f_x}$$

where t_0 is the width of the pulse packet at half voltage, T is the time between packets, and f_p is the frequency of the waves making up the packets. The distance between half-power points on the above frequency spectrum is $0.61 f_p$. The over-all bandwidth of the tank and amplifier should be a small amount greater than this to obtain the desired packet width at the detector, when the input of the tank is driven with a pulse having relatively rapid rise and slow fall.

Under the conditions just described, the detector output will drop to about 15 per cent of its maximum value between two pulses which are separated by a space, but will not drop at all (in fact, will rise slightly) between two adjacent pulses. Such a situation is somewhat unconventional in a pulse transmission system, but it will be recalled that the output pulses are not permitted to re-enter the line, but are only used to gate timing pulses. Therefore, if the latter are themselves sharp enough, it is only necessary for the output signal to be above or below the critical gating level by a safe amount in order to pas or reject a timing pulse. The gating signal can therefor if necessary, retain its full amplitude for a number pulse times if it represents an unbroken sequence pulses, only dropping below the critical level when it necessary to reject one or more pulses. Such a represe tation of pulse signals is known as a "pulse envelope system, and can be shown to result in a reduction by factor of two in the bandwidth requirements for trans fer of a given amount of information per unit time.

IV. Construction

Further details of the mercury tanks are shown i Figs. 2, 3, and 4. Fig. 2 shows the construction of th

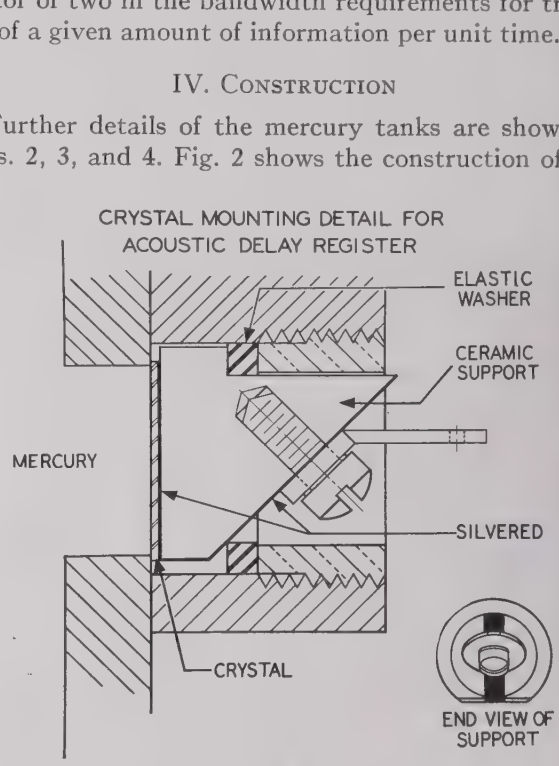

Fig. 2

crystal mounting. The ceramic support is silvered; on surface of the crystal is silvered, and the two are soldere together. The silvering is carried around the support t provide a connection to the back electrode; the mercur itself forms the front electrode. The support provide the proper acoustic termination for waves from the rea surface of the crystal. A threaded brass retaining rin and elastic washer hold the crystal support in the en of the mercury tank. Fig. 3 shows three crystals mounte on their steatite supports.

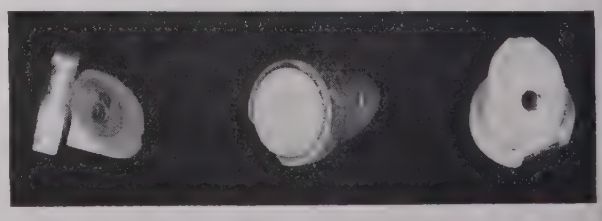

Fig. 3

Fig. 4 shows the various components of an 18-chan nel tank. The end cells into which the crystal mounting

[7] See sections, 2, 5, 3, and 5, respectively, of footnote references 3–5.

[8] G. A. Campbell and R. M. Foster, "Fourier Integrals for Practical Application," Bell System Monograph B-584; 1931.

are shown at bottom left and right, surrounded by ounted crystal, retaining bushings, and short coaxial nnectors which are used between crystal element and circulation chassis. The mounted crystals fit into the les in the end cells and bear against retaining plates aving holes slightly smaller than the crystals; these ates are shown just below the tank cylinder, with the ers used to reduce cross talk. Above the cylinder are e rods which hold the assembly together, and at the ght is one of the supports which center the assembled nk inside the outer jacket. The components at bottom nter form an expansion chamber for the mercury. All etal parts in contact with mercury are made of a spe-al stainless steel.

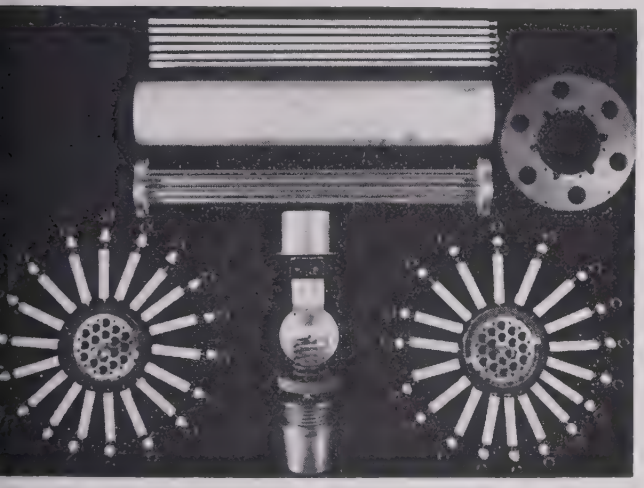

Fig. 4

The outer surface of the cylinder will be coated with sulating resin and wound with resistance wire, which rves as a heating element. The latter is part of a tem-erature control system which keeps the repetition rate the pulses in synchronism with the clock or timing ulses. Here we note another advantage of the common ercury pool over a group of separate tanks: it is now sy to keep the temperature of all channels uniform. In e proposed designs of some previous computers, com-ensation for changes in mercury temperature was made y varying the clock frequency through a reactance ibe. Now, because of the ease with which the mercury mperature can be controlled, it becomes possible to vise this procedure and use a crystal-controlled oscil-tor in the clock, which in turn controls the mercury mperature, and hence controls the circulation time of e tanks. This greatly simplified the problem of keeping l parts of the computer in synchronism, and in particu-r makes it easy to use a single clock to control two mputers. The latter arrangement is valuable in cases here it is necessary to provide continuous checking of sults by intercomparing the operation of the two com-uters.

Fig. 5 is a block diagram of the temperature control system. One channel in the multichannel tank is reserved as a temperature control channel. By means of a frequency divider, the timing pulse frequency is divided down to produce a series of pulses having a repetition rate corresponding to an integral submultiple of the total number of pulses stored in the tank. This submultiple is ordinarily made equal to the number of pulses in each number stored in a tank; for example, if each num-

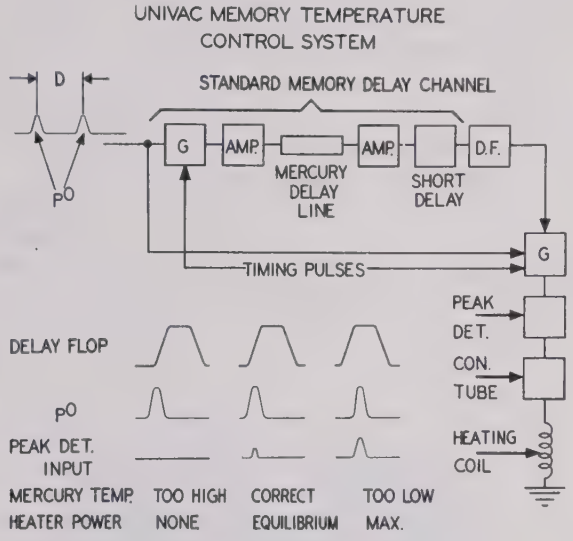

Fig. 5

ber requires 50 pulses to represent it, the basic pulse rate is divided by 50. Division is accomplished by means of an electric delay line with a single pulse circulating in it. This technique yields a large division ratio of precision equal to that obtained by a successive division, using a number of multivitiators or blocking oscillators in cascade. The pulses from the divider, which we may think of as corresponding to the last pulse in each number, are referred to as $p0$ pulses. They are supplied to the input gate of the temperature control channel and also, through the mercury line and a delay flop, to the output coincidence gate. Timing pulses are also applied to these gates to increase the accuracy of temperature control. This technique results in an over-all temperature control of $\pm 0.3°$ absolute.

If the capacity of a memory channel is, say, twenty 12-digit numbers, and the delay is accurately adjusted, then any $p0$ pulse entering the input will, after going through the line, coincide with the twentieth subsequent $p0$ pulse at the output gate. If the temperature is a little too high, however, the pulse through the line will arrive too late and fail to gate the twentieth following pulse. If the temperature is correct, the pulse from the line will have just reached half amplitude when the timing pulse, as gated by the twentieth following pulse, arrives, and the gate will produce an output pulse of intermediate amplitude. Finally, if the line is too cold, the pulse through the line will arrive too soon; ordinarily it would

again fail to gate the twentieth following pulse, but if it is used to trigger a delay flop, the effect is to produce a pulse of maximum amplitude at the output of the gate. Thus the effect of the circuits shown is to produce an output signal whose amplitude is proportional to the amount of heat which should be supplied to the line to assure perfect synchronism. The gate output is fed to a peak detector to smooth out the pulse signal into a continuous signal which, after suitable amplification, controls the current to the heater coil around the tank.

A similar effect could have been achieved by using an average detector whose output is proportional to the amount of overlap of the $p0$ pulse and the signal from the line. Because of the low duty cycle, however, the detector output would be small and would require a considerable amount of amplification. The circuit used, giving an output whose amplitude varies in accordance with the relative position of a standard pulse and the relatively gradually sloping front of the signal from the line, in combination with the peak detector, gives a signal which requires little amplification.

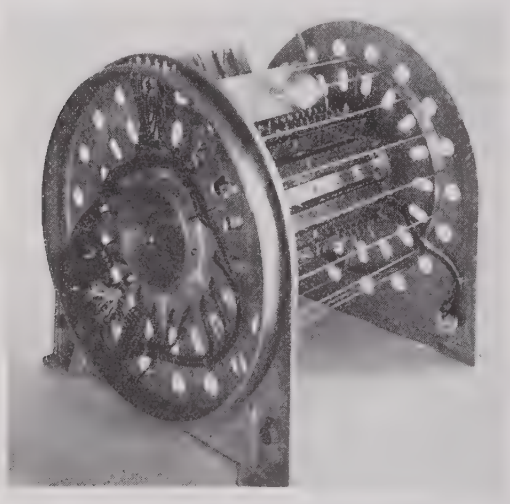

Fig. 6

Fig. 6 shows the outer cylinder mounted between its supporting end castings and surrounded by terminal boards to receive the plug-in recirculation chassis. The space between inner and outer cylinders is filled with santocel, a good thermal insulator. The radially disposed

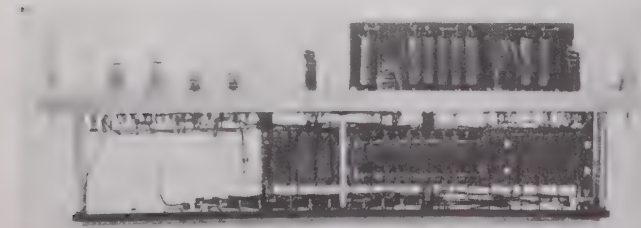

Fig. 7

holes near the ends of the cylinder receive the short c axial connectors shown in Fig. 4.

In the plug-in recirculation chassis shown in Fig. the four tubes at the left are the band-pass amplifier type 6AK5. These are followed by a detector using ge manium crystals, and an AVC circuit. The latter fe ture is essential to compensate for changes in gain du to tube aging. A 50C5 amplifies the detector output si nal and applies it to the output and recirculation gate the recirculation and input gates feed another 50C which drives the clock gate. Timing pulses from the ma ter oscillator are fed to the other input of this gat Finally, the output of the clock gate is fed to a 50C driver which applies the reshaped and retimed signals t the crystal at the input end of the mercury column.

All gating is done by means of germanium diode which, because of their low forward impedance and sma shunt capacity, are useful for gating operations at hig pulse rates. The gating circuits are shown in Fig.

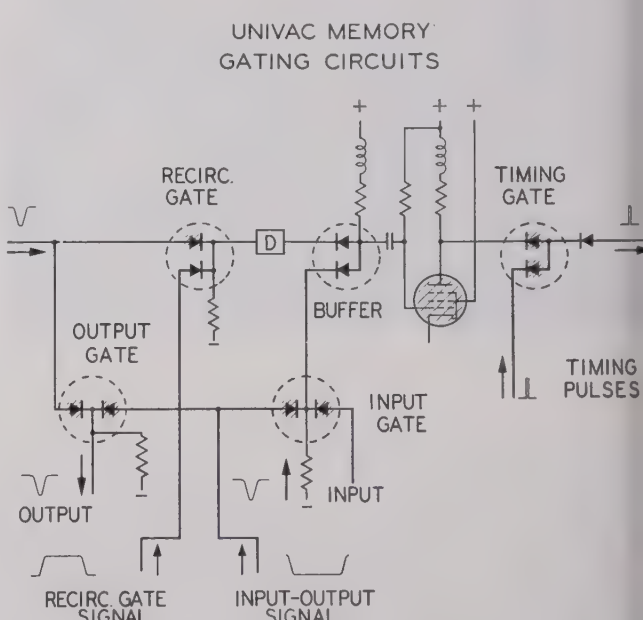

Fig. 8

Crystals which are normally conducting are shaded. Th load resistors and the voltages to which they are co nected are so chosen that the signal developed across th load resistor will not be sufficient to actuate the follow ing circuit if only one of the crystals is conducting. How ever, it will actuate the following circuits if both crystal do not conduct. For example, an input signal will cut o current in the right-hand crystal of the input gate an cause somewhat less voltage drop across the load re sistor, but the resulting signal will be of insufficient am plitude to cause the lower buffer crystal to conduct, un less a gate signal is simultaneously applied to the left hand crystal of the input gate. The gates are not ampli tude-sensitive in the usual sense, as input signals nee

ly have sufficient amplitude to cut off their respective ystals.

Wave forms of the signals at various points in the stem are shown in Fig. 9. The top curve shows the out-

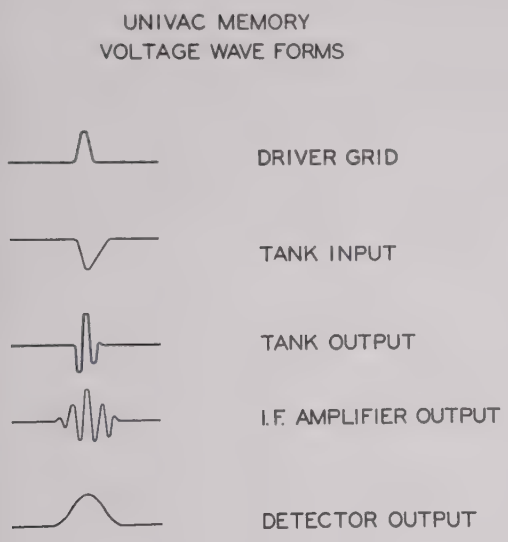

Fig. 9

ıt of the clock gate as it is applied to the grid of the iver tube. Below is the voltage developed by the out-ıt crystal; it goes through about one and one-half os-llations as a result of bandwidth limitations in the nk.

The next curve is the output signal from the band-ass amplifier. In order to obtain adequate gain without o many stages, only sufficient bandwidth to pass a sable signal has been used; hence the amplifier intro-uces a noticeable amount of phase and amplitude dis-rtion. The lower curve shows the detector output volt-ge; it is this signal which is amplified and used to gate a ming pulse, resulting in a signal of the form shown in ıe top curve in Fig. 9.

V. Conclusion

This paper has reviewed the results of developmental study of the basic acoustic delay line memory. As a result of this development, it was shown that it is possible to transmit and receive intelligence at the rate of 5,000,000 binary digits per second. This is a considerable extension over previous telegraphic keying rates and is thus of interest in its relation to information theory.

Of the various techniques employed to extend the useful frequency range of memory systems, the application of the pulse envelope system is probably the most significant in that it represents an important contribution to the concept of transmission of intelligence. Its importance lies in the fact that it provides a simple means of increasing the effective bandwidth of a transmission system without the necessity for a corresponding improvement in the characteristics of the components of the system. Another contribution was the application of germanium crystal gating circuits which permitted the effective control of signals at this high pulse repetition rate. The circuits developed in this work have found many other applications in the arithmetic and control circuits of computers.

The use of a single pool of mercury for multiple channels of information reduced the physical size of the whole memory system. As a result of the intimacy of metal contact between the channels, temperature gradients were greatly reduced, the temperature control problem was simplified, and synchronous operation of two computers for checking purposes facilitated.

The acoustic memory described represents one of the first practical applications of techniques which have extended the useful audio transmission range from its previous limits of some 150,000 cps to a new present value of 30 million cps or more.

This paper has reviewed the several improvements that have been made which helped to extend the useful acoustic transmission range.

CORRECTION

A drafting error in the paper, "A Digital Computer for Scientific Applications," by C. F. West and J. E. DeTurk, which appeared on pages 1452–1460 in the December, 1948, issue of the Proceedings of the I.R.E., has been brought to the attention of the editors by the authors. The error was made in Fig. 10, on page 1457, in which the signals feeding through 1N34 crystals to the grids of the indicated flip-flops should be transposed.

An Analysis of Magnetic Amplifiers with Feedback*

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D. C. BEAUMARIAGE†, STUDENT MEMBER, IRE

Summary—**Two ways of obtaining feedback in magnetic amplifiers are studied. One, termed external feedback, employs a bridge rectifier and separate feedback coils, while the other, termed self feedback, uses two rectifiers so disposed that separate feedback windings are unnecessary. Commutation of the rectifiers, apparently not considered heretofore, is shown to be of major importance. The calculated currents are confirmed satisfactorily, both in shape and magnitude, by experimental checks. The two feedback methods are compared, and it is found that self feedback is advantageous.**

Introduction

A SIMPLE magnetic amplifier consists of two identical single-phase transformers having one pair of similar windings in series with an ac source and a load, and the other pair in series with a dc source. A small direct current in the latter windings, through saturation of the cores, determines the inductive reactance presented on the ac side and so controls the load power. This and similar types of magnetic amplifiers are treated in a companion paper.[1] A more complex type achieves greater sensitivity through positive feedback or self-excitation schemes well known in practice.[2–6] The purpose here is to treat these feedback circuits more precisely than has been done before.

One way of obtaining feedback is shown in Fig. 5, which is derived from Fig. 1 of the companion paper[1] by adding a third set of windings N_3 to each core and connecting them to a bridge rectifier in series with the load. The mmf's produced in the feedback windings N_3 reinforce those produced in the control windings N_1, so that a given average saturation is achieved with less control current than would be needed without feedback. This arrangement is identified here by the term "external feedback."

A less obvious way of obtaining feedback is shown in Fig. 1, which is identical with Fig. 4 of the companion paper,[1] except for the addition of a rectifier in series with each of the two load coils. During one half cycle the load current links one core and during the next half cycle the other, thus producing direct components of mmf which aid those produced by the control current and so gives feedback without additional windings. The term "self feedback" is used for this circuit.

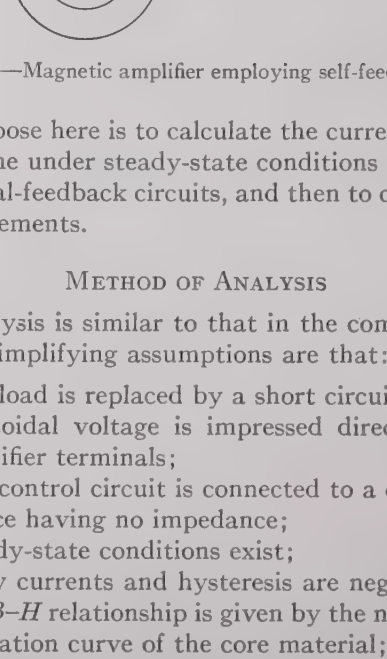

Fig. 1—Magnetic amplifier employing self-feedback.

The purpose here is to calculate the currents as functions of time under steady-state conditions for the self- and external-feedback circuits, and then to compare the two arrangements.

Method of Analysis

The analysis is similar to that in the companion paper.[1] The simplifying assumptions are that:

(a) The load is replaced by a short circuit, so that a sinusoidal voltage is impressed directly on the amplifier terminals;
(b) The control circuit is connected to a constant dc source having no impedance;
(c) Steady-state conditions exist;
(d) Eddy currents and hysteresis are negligible, and the B–H relationship is given by the normal magnetization curve of the core material;
(e) Leakage fluxes are negligible, and the flux density is uniform throughout each core;
(f) Coil resistances are very small, although not neglected entirely; and
(g) Rectifier reverse resistances are infinite, and forward resistances are very small and constant.

Briefly, the procedure is to apply Kirchhoff's voltage law to the electric network and solve for the magnetic fluxes in the cores; use the magnetization curve of the material to find the corresponding mmf's; apply Ampere's law to each core to relate mmf's and currents;

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[1] D. W. VerPlanck and M. Fishman, "An analysis of interlinked electric and magnetic networks with application to magnetic amplifiers," scheduled to appear in an early issue of the PROC. I.R.E.

[2] E. C. Wentz, "Direct-current controlled reactor," *Electric Jour.*, vol. 28, p. 561; October, 1931.

[3] T. Buchhold, "Über gleichstromvormagnetisierte Wechselstromdrosselspulen und deren Rückkopplung," *Arch. für Elek.*, vol. 36, p. 221; month, 1942.

[4] T. Buchhold, "Zur Theorie des magnetischen Verstärkers," *Arch. fur Elek.*, vol. 37, p. 197; April, 1943.

[5] H. S. Kirschbaum and E. L. Harder, "A balanced amplifier using biased saturable core reactors," *Trans. AIEE*, (*Elec. Eng.*) vol. 66, pp. 273–278; 1947.

[6] A. Uno Lamm, "Some fundamentals of a theory of the transductor or magnetic amplifier," *Trans. AIEE*, (*Elec. Eng.*) vol. 66, pp. 1078–1085; 1947.

, finally, solve for the currents. In doing this, it is essary to account for commutation periods in which rectifiers conduct simultaneously.

s it is the simpler of the two arrangements to be lied, the self-feedback circuit is treated first.

Self-Feedback Circuit

eferring to Fig. 1, and replacing the load by a short uit, Kirchhoff's voltage law gives three equations resented in matrix form as

$$\begin{Vmatrix} E_{dc} \\ {}_m \cos \omega t \\ {}_m \cos \omega t \end{Vmatrix} = \begin{Vmatrix} N_1 & N_1 \\ N_2 & 0 \\ 0 & -N_2 \end{Vmatrix} \cdot \begin{Vmatrix} \dot{\phi}_\alpha \\ \dot{\phi}_\beta \end{Vmatrix} + \begin{Vmatrix} 2R_1 & 0 & 0 \\ 0 & R_2 + R_r & 0 \\ 0 & 0 & R_2 + R_r \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix}. \quad (1)$$

s units are used and the symbols are as shown in 1, with the addition that R_1 and R_2 are, respecly, the resistances of each of the single windings N_1 N_2, and R_r is the rectifier resistance equal either to mall and constant value or infinity, depending on ther or not the rectifier is conducting.

When rectifier a alone is conducting, the *positive pe-*, only the first two equations of (1) are useful. Simy, in the *negative period*, when b alone conducts, the first and last of (1) apply. All three are useful he *commutation period* when both a and b conduct. he fluxes are found by simultaneous solution of the ropriate voltage equations, assuming that the reance drops (and hence E_{dc}) are negligible. Flux den- B_α and B_β in the respective cores are then found dividing the fluxes by the core area A. Each period t be treated separately but in each case the result ne same, namely

$$B_\alpha = B_m \sin \omega t + B_0 \quad (2)$$

$$B_\beta = -B_m \sin \omega t + B_0 \quad (3)$$

re $B_m = E_m/N_2\omega A$. B_0, the average flux density in core, to be determined, is the same throughout the le because the linkage of the cores by closed circuits inite resistance prevents sudden changes of flux at transitions between periods; also, considering symry, it is clear that B_0 is the same for each core. For commutation period, only two of the three voltage ations are needed to determine the flux densities; third equation is useful later.

he field intensities are derived from the flux densities ough graphical use of the magnetization curve, as in . 5 of the companion paper.[1] In functional notation, relationship is

$$H_\alpha = H(B_m \sin \omega t + B_0) \quad (4)$$

$$H_\beta = H(-B_m \sin \omega t + B_0). \quad (5)$$

The required mmf's are, respectively, LH_α and LH_β where L is the mean length of core. Then, applying Ampere's law, two equations result which in matrix form are:

$$\begin{Vmatrix} N_1 & N_2 & 0 \\ N_1 & 0 & -N_2 \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix} = \begin{Vmatrix} LH_\alpha \\ LH_\beta \end{Vmatrix}. \quad (6)$$

In the positive and negative periods, i_3 and i_2 are respectively equal to zero, and the remaining two currents are found directly as shown in Table I.

TABLE I

Expressions for Currents in Amplifier with Self-Feedback

Period	Currents
Positive	$i_1 = LH_\beta/N_1$ $i_2 = L(H_\alpha - H_\beta)/N_2$ $i_3 = 0$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$
Commutation	$i_1 = [k'L(H_\alpha + H_\beta)/2 + N_1 I_1]/N_1(1+k')$ $i_2 = [k'L(H_\alpha - H_\beta)/2 + LH_\alpha - N_1 I_1]/N_2(1+k')$ $i_3 = [k'L(H_\alpha - H_\beta)/2 - LH_\beta + N_1 I_1]/N_2(1+k')$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$
Negative	$i_1 = LH_\alpha/N_1$ $i_2 = 0$ $i_3 = L(H_\alpha - H_\beta)/N_2$ $i_2 + i_3 = L(H_\alpha - H_\beta)/N_2$

In the commutation period all three currents have values, and so an additional equation is needed. This is obtained by eliminating the rates of change of flux from (1) but without neglecting the resistance drops. This equation and (6) are solved simultaneously, with the results shown in Table I. In these expressions,

$$I_1 = \frac{E_{dc}}{2R_1} = \frac{1}{2\pi} \int_0^{2\pi} i_1 d(\omega t), \quad (7)$$

the average (dc) value of the control current, and

$$k' = \left(\frac{N_1}{N_2}\right)^2 \frac{R_2 + R_r}{R_1}. \quad (8)$$

Thus, as in circuits with parallel-connected load coils,[1] resistances determine the currents when the mmf can be supplied in part by currents in either of two paths, even though the resistances may be negligible in other respects. Here the two paths are the control circuit and that through the two load coils and two rectifiers in series in the forward direction.

Although expressions for the currents have been derived for all periods of the cycle, the average flux density B_0, which enters all of the expressions, and the times of transition between periods have not been found. To determine these unknowns requires a graphical trial and error process, illustrated in Fig. 2, and consisting of the following steps:

(a) A value of B_0 is assumed.

(b) The centers of the commutation periods are located by plotting $N_2(i_2+i_3)$, which, as is indicated in Table I, is independent of transition times.

(c) As a first trial, commutation is assumed instantaneous and N_1i_1 is plotted using the expressions in Table I for the positive and negative periods (positive and negative referring to the sign of i_2+i_3) and continuing the plots to their intersections as shown by (1) in Fig. 2.

(d) The average control excitation N_1I_1, defined by (7), is found from the area under N_1i_1 as plotted in (c).

(e) The value of N_1I_1 from (d) is used to obtain a second approximation to N_1i_1 in the commutation period, (2) in Fig. 2.

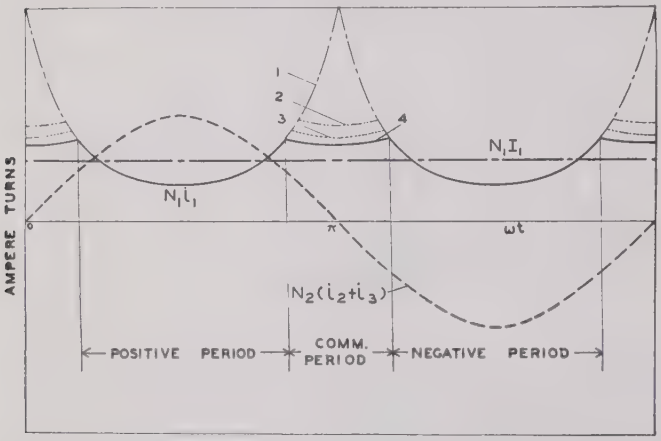

Fig. 2—Successive approximations to N_1i_1 and the limits of the commutation period. The first three approximations to N_1i_1 and the correct value are shown by 1, 2, 3 and 4, respectively.

$N_1=$ 500 turns	$\omega=2\pi 60$ sec^{-1}	$R_1=$ 35 ohms
$N_2=4000$ turns	$L=0.18$ meters	$R_2=190$ ohms
$E_m=140\sqrt{2}$ volts	$A=4.2\times10^{-4}$ (meters)^2	$R_r=$ 50 ohms

This leads to a new value of N_1I_1, which is used to correct N_1i_1, and so on. The process converges rapidly, the fourth approximation to N_1I_1 in Fig. 2 being essentially correct. Thus the transition times are located, and the currents are completely determined for the assumed B_0. If the problem must be solved for a specified I_1, a succession of values of B_0 are tried until the one corresponding to the desired I_1 is found.

Experimental checks are given in Figs. 3 and 4, which are for the circuit parameters of Fig. 2 and for the magnetization curve shown in the companion paper.[1] Fig. 3 compares calculated currents (plotted as ampere-turns) with points measured from oscillograms for the same conditions as Fig. 2. Fig. 4 compares calculated and measured external characteristics over a range of N_1I_1 with other quantities fixed at the values of Figs. 2 and 3. The load current i_2+i_3 was measured with a rectifier-type ammeter, and calculated from areas under curves of instantaneous currents like that in Fig. 3.

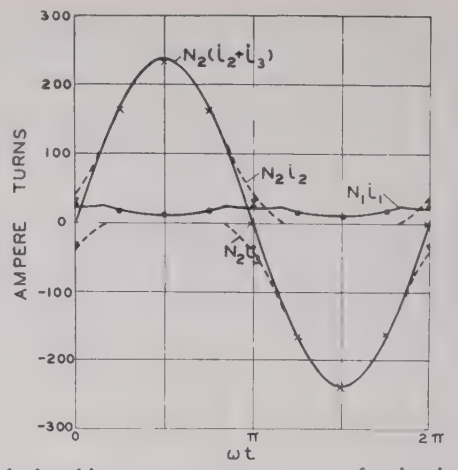

Fig. 3—Calculated instantaneous ampere turns for the circuit of Fig 1 compared with points measured from oscillograms. Circuit quantities as in Fig. 2 and $N_1I_1=17$.

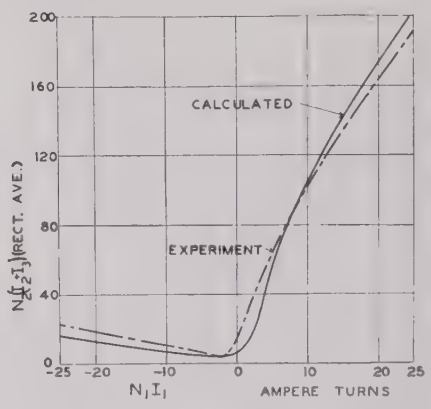

Fig. 4—Comparison of calculated and experimentally determined characteristic curves for the self-feedback circuit, Fig. 1. Circuit quantities as given for Fig. 2, except that N_1I_1 is varied.

External-Feedback Circuit

In the circuit with external feedback, Fig. 5, again one must consider three periods: the positive period

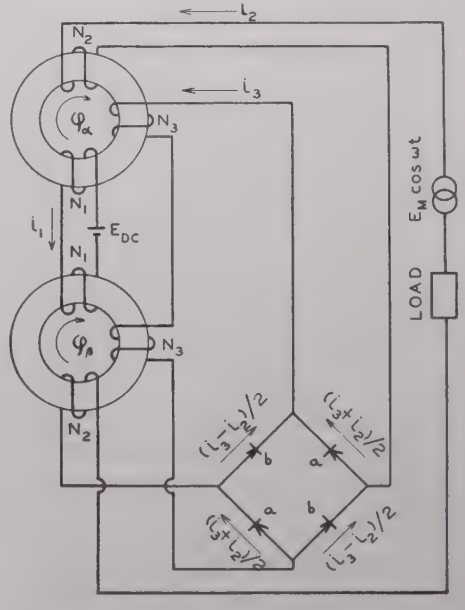

Fig. 5—Magnetic amplifier employing external feedback.

hen rectifiers *aa* alone conduct; the negative, when conduct; and commutation, when all four conduct. For the control circuit one voltage equation applies roughout, namely,

$$E_{dc} = N_1(\dot{\phi}_\alpha + \dot{\phi}_\beta) + 2R_1 i_1. \tag{9}$$

During the positive period, $i_3 = i_2$, and the voltage uation for the load circuit is

$$E_m \cos \omega t = (N_3 + N_2)\dot{\phi}_\alpha + (N_3 - N_2)\dot{\phi}_\beta + 2(R_2 + R_3 + R_r)i_2. \tag{10}$$

the negative period, $i_3 = -i_2$, and the voltage equation may be obtained from (10) by changing N_3 to $-N_3$. For the commutation period, (10) is replaced by two ltage equations: one for the loop through the generor, the coils N_2 and either pair of rectifiers *ab*,

$$E_m \cos \omega t = N_2(\dot{\phi}_\alpha - \dot{\phi}_\beta) + (2R_2 + R_r)i_2, \tag{11}$$

d the other for the loop through the coils N_3 and a ir of rectifiers *ab*,

$$0 = N_3(\dot{\phi}_\alpha + \dot{\phi}_\beta) + (2R_3 + R_r)i_3. \tag{12}$$

Simultaneous solution of the appropriate voltage uations for each of the periods results in expressions r the flux densities B_α and B_β in the respective cores entical with (2) and (3) for the self-feedback case, expt that here $B_m = E_m/2N_2\omega A$. As in the previous case e field intensities are given by (4) and (5).

The mmf conditions in matrix form are:

$$\begin{Vmatrix} N_1 & N_2 & N_3 \\ N_1 & -N_2 & N_3 \end{Vmatrix} \cdot \begin{Vmatrix} i_1 \\ i_2 \\ i_3 \end{Vmatrix} = \begin{Vmatrix} LH_\alpha \\ LH_\beta \end{Vmatrix}. \tag{13}$$

or the positive and negative periods, i_3 is equal respecvely to $+i_2$ and $-i_2$, and (13) may be solved for the rrents. For the commutation period, it is necessary get another equation from the voltage conditions, as the self-feedback case. Doing this, the currents in all ree periods are found readily, as shown in Table II. ere I_1 is defined exactly as in (7), and

$$k'' = \left(\frac{N_1}{N_3}\right)^2 \frac{2R_3 + R_r}{2R_1}. \tag{14}$$

TABLE II

Expressions for Currents in Amplifier with External Feedback

Period	Currents
Positive	$i_1 = L[(1 - N_3/N_2)H_\alpha + (1 + N_3/N_2)H_\beta]/2N_1$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = i_2$
Commutation	$i_1 = [k''L(H_\alpha + H_\beta)/2 + N_1 I_1]/N_1(1 + k'')$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = [L(H_\alpha + H_\beta)/2 - N_1 I_1]/N_3(1 + k'')$
Negative	$i_1 = L[(1 + N_3/N_2)H_\alpha + (1 - N_3/N_2)H_\beta]/2N_1$ $i_2 = L[H_\alpha - H_\beta]/2N_2$ $i_3 = -i_2$

The limits of the commutation period and the relation between B_0 and I_1 are found by trial, as explained for the self-feedback case.

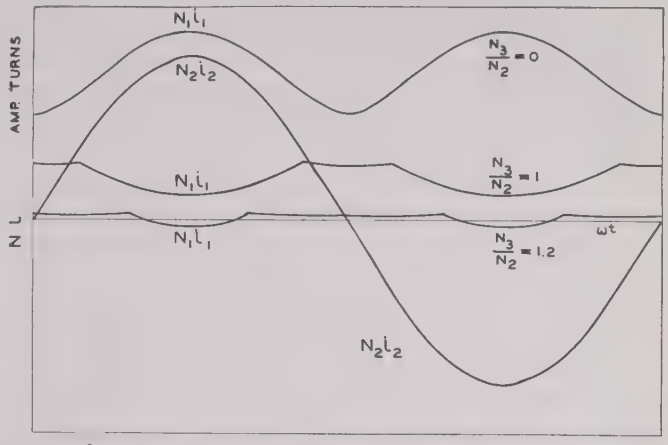

Fig. 6—Calculated excitation curves for the circuit of Fig. 5, showing the effect of varying the number of feedback turns while keeping the total amount of copper in feedback and control windings constant.

Fig. 6 and Table III show an application of the foregoing analysis of the magnetic amplifier with external feedback. The purpose is to show how varying the feedback turns affects the control excitation and power. The load windings N_2, the magnetic quantities B_m and B_0, the ac voltage, and the load current remain the same throughout. That is, conditions on the ac side are identical for the three cases shown. The total amount of copper in the control and feedback windings together is kept fixed by holding the quantity $(N_1^2/R_1 + N_3^2/R_3)$ constant, but the distribution of copper between these windings is varied. Forward resistances of the rectifiers are neglected in these calculations. The last column of Table III shows how greatly the power which must be supplied by the control source can be reduced in this particular case by using feedback. These results, however, must be viewed with some caution, because another effect of high feedback may be to cause the amplifier to become unstable. The present analysis is also useful in the study of stability.

TABLE III

Effect of Feedback Turn Ratio on Control Power Required for a Particular Case

N_3/N_2	N_1^2/R_1	N_3^2/R_3	$N_1 I_1$	$I_1^2 R_1 = (N_1 I_1)^2/(N_1^2/R_1)$
0	18,400	0	85	0.39
1.0	3,400	15,000	25	0.18
1.2	400	18,000	1.4	0.0049

Comparison of Self- and External Feedback

In comparing the self- and external-feedback circuits, Figs. 1 and 5, one must note that the latter is more flexible in that there may be various amounts of feedback, depending on the number of turns N_3. The comparison, therefore, is made on the basis that the feed-

back and load coils in the external-feedback circuit are equal; that is,

$$N_3 = N_2 \tag{15}$$

$$R_3 = R_2. \tag{16}$$

With this condition, Tables I and II show that the expressions for control currents in the two cases and for the load currents (i_2 and i_2+i_3, respectively) become identical if

$$k' = k'' \tag{17}$$

$$N_1' = N_1'' \tag{18}$$

$$N_2' = 2N_2'' \tag{19}$$

where single primes are used for self-feedback quantities and double primes for external feedback. The currents actually will be identical if L, H_α, and H_β are the same in the two cases. This identity will exist if core dimensions and material are identical, and impressed ac voltages and average values of control current I_1 are the same.

Having established these conditions for identity of magnetic conditions and of external currents in the two cases, it follows that

$$R_2' + R_r' = 2(2R_2'' + R_r''), \tag{20}$$

provided that

$$R_1' = R_1''. \tag{21}$$

With these relations between resistances and the other conditions of equivalence, it is found, using the expressions for the currents in Tables I and II, that the i^2R losses in the control windings are the same at every instant, and similarly in the load and feedback windings and the rectifiers taken together. Thus the performance of the two arrangements will be identical in all respects under the conditions set. Examination of the relations between resistances and numbers of turns shows, however, that external feedback requires twice as much copper in the load and feedback coils together as is required in the load coils of the self-feedback circuit.

Thus, for equivalent performance, the self-feedback arrangement, Fig. 1, can be made smaller than that employing external feedback, Fig. 5, and is therefore superior provided that the feedback turn ratio N_3/N_2 desired is unity. For greater feedback turn ratio, a combination of the arrangements is indicated;[6] rectifiers being provided as in Fig. 1 to give most of the feedback effect, and booster windings like N_3 in Fig. 5 being added to supply additional feedback as desired.

Conclusion

Two well-known methods for obtaining feedback in magnetic amplifiers have been analyzed quantitatively, with results that are checked closely by experiment. The analysis has been used to show quantitatively how feedback reduces the control excitation requirements and to compare the two feedback arrangements.

The analysis takes account of commutation, or simultaneous conduction of the rectifiers, a phenomenon which does not seem to have been considered so far, although it may have been observed.[3]

Electrical Network Analyzers for the Solution of Electromagnetic Field Problems*

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Part II
OPERATION

Summary—**This is the second part of a paper on the design, construction, testing, and operation of electrical networks capable of obtaining solutions of the wave equation in both two-dimensional, axially symmetric, cylindrical co-ordinates and rectangular co-ordinates. The circuit yields information on *TEM*, $TM_{om}(n)$ and $TE_{om}(n)$ modes of concentric lines, waveguides, and resonators. Field plots of electric- and magnetic-field configurations are directly available, and resonant frequencies of cavities, equivalent impedance, reactance characteristics, Q, and propagation characteristics can be determined.**

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Introduction

IN PART I, the general theory, design, and construction of two network analyzers for solution of the wave equation in both two-dimensional, axial symmetric, cylindrical co-ordinates and rectangular co-ordinates were discussed. This section will deal only with

testing and general use of the networks. These net-rk analyzers have been used at Stanford University about two years, and have been found to be very pful in the general solutions of field problems. The olicability of the network has been well established comparing numerous results with known cases. The ious boundary conditions that must be met and their parent position in respect to the network are dis-sed. Techniques for testing and solving special prob-ns, such as resonator characteristics, obtaining of d plots, and the measurement of shunt resistances d Q, are presented.

IV. Network Adjustment

Tuning the Cylindrical-Co-ordinate Network

The only tuning involved on the network is that of usting the trimmer capacitors. The inductances can assumed to be correct from their original calibra-n. In tuning the capacitor, it was found that the most urate method of adjusting the capacitors was a onance-substitution method applied to each ca-itor individually. Since the capacitance values vary h radius, this involves a number of different adjust-nts consuming considerable time. A careful adjust-nt of all the network capacitance values requires ap-ximately two man days of work.

Tuning the Rectangular-Co-ordinate Network

Similarly with this network, the only adjustment is t of the capacitance values. The uniformity of his work makes these operations relatively simple.

Methods were developed for checking both continuity coils and adjusting capacitor values which did not re-re that any coils be unplugged from their operating itions in the clips.

Adjustment of capacitance values was facilitated by fact that the coils uniformly had a natural resonant quency of 2 Mc. At this frequency, they exhibited a y high resistance compared to the proper value of acitive reactance. Accordingly, a special coil was igned to be clipped across the capacitors and to be onated with the proper value of capacitance at 2 Mc. sonance was indicated with a Measurements Corpo-ion Megacycle Meter. By this method, it was pos-le to adjust all the capacitances to a high degree of uracy in approximately one-half man day.

V. Use of the Networks

Use of the Cylindrical Network

1. Determination of Cavity Resonant Frequencies. The termination of the resonant frequencies of cavity onators is relatively simple. A scaled outline of the vity is set up on the network by opening the circuit the boundary corresponding to the cavity walls for M_0 modes and shorting capacitors to ground for TE_0 des. The network is then excited from a variable-fre-ency oscillator, and the frequency of oscillation varied til a resonance is obtained. Excitation of the network is achieved by simply connecting the ground plane and an arbitrary point of the network lattice to the oscil-lator through a high series resistance (100,000 ohms) simulating very loose coupling. Coupling may also be had inductively by a coupling loop in the vicinity of one of the network coils. It should be borne in mind that, for a given boundary condition, i.e., either open or short circuit there will be revealed resonances of the cor-responding mode type only; i.e., TM_0 and TE_0, respec-tively. It should also be borne in mind that the cor-responding edge of the field is different for the two types of boundary conditions. When the conductor is represented by an open circuit, the effective field bound-ary is very nearly one-half section beyond the capacitor at which the network is opened. For the short-circuit type of boundary, however, the effective edge of the field is at the short. These end corrections are the same by test for either radial or longitudinal boundaries with the cylindrical-co-ordinate network. A similar end cor-rection applies to the rectangular-co-ordinate network.

2. Resonator Tuning Curves. Resonator tuning curves as one dimension of a cavity is varied are obtained by the method related in the previous section. It is only necessary to record and plot resonant frequency as a function of the corresponding cavity dimension. A curve of resonant wavelength versus cavity length for a concentric-line resonator is shown in Fig. 6. This figure exhibits several groups of curves for different modes and different ratios of radii in the concentric-line por-tion. The curves have been extended to allow the plunger to protrude into the end space of the resonator.

3. Cylindrical Cavity Field Plots. For cylindrical TEM and TM_0 modes, the modes of usual interest, complete magnetic- and electric-field plots are obtained from the junction voltage-to-ground data at network resonance. Since voltage to ground corresponds to magnetic field

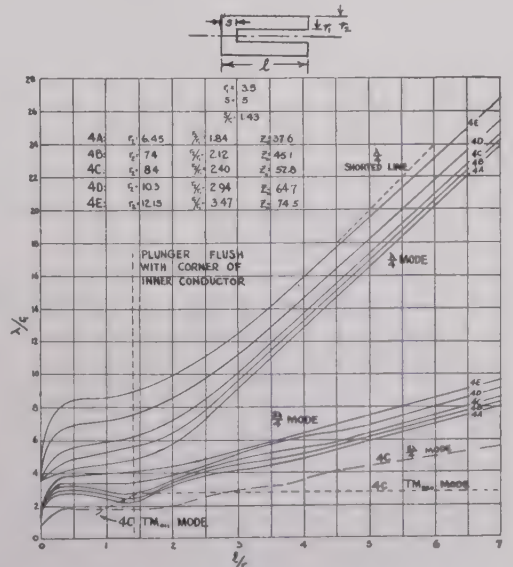

Fig. 6—Resonant wavelength of a loaded concentric-line resonator as a function of length.

multiplied by radius, it is only necessary to divide voltage values by the radius to get the relative magnetic field.

A type of electric-field plot is given directly by contours of constant network voltage to ground. Such contours have the direction of electric flux lines. Flux density is, however, given by line density of the voltage contours divided by radius. While this has not been a common convention for plotting electric flux fields, it should be pointed out that it is the only correct convention for cylindrical geometries which permits the use of continuous lines when such lines are not all parallel to the axis. In working with the network it has been found desirable to use the above convention exclusively because, in the first place, it is the only consistent convention which can be used for general field configurations, and, in the second place, such field plots are given directly by contours of constant voltage on the network.

The reason for the above relations is seen from equations (7) and (8) in Part I. Here it is seen that the electric flux density multiplied by radius is a vector whose magnitude is proportional to the gradient of rH_ϕ (hence of network voltage to ground), and whose direction is normal to the gradient of rH_ϕ (or network voltage). Since the direction of the electric flux is normal to the gradient of rH_ϕ (network voltage), which, in turn, is normal to the contours rH_ϕ (network voltage), then the electric flux has the direction of contours of constant rH_ϕ (network voltage). Since the gradient of rH_ϕ (network voltage) is proportional to electric field multiplied by radius, then the electric flux density is given by the density of lines of constant rH_ϕ (or network voltage) divided by the radius.

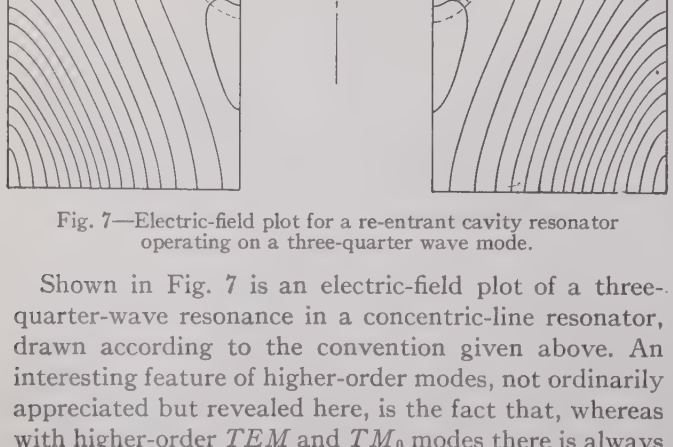

Fig. 7—Electric-field plot for a re-entrant cavity resonator operating on a three-quarter wave mode.

Shown in Fig. 7 is an electric-field plot of a three-quarter-wave resonance in a concentric-line resonator, drawn according to the convention given above. An interesting feature of higher-order modes, not ordinarily appreciated but revealed here, is the fact that, whereas with higher-order TEM and TM_0 modes there is always a nodal surface of magnetic field, there need not be a true nodal surface of electric field. There will, however, always be a pseudo-nodal surface of electric field which can be drawn through a point of zero electric field, which will always exist in such a way that no power is transmitted across this surface and such that, if the surface were replaced by a conductor, the two resonator thus created would have field configurations like por tions of the orginal and would resonate at the same frequency as the entire cavity. The "nodal" surface o electric field having these properties is shown by a dashed line in Fig. 7.

4. Determination of Shunt Resistance and Q of Cylindrical Cavity Resonators. The cylindrical-co-ordinate network can be used to measure the shunt resistance and Q of cavity resonators directly. The details of the method of measurement and a discussion of some results obtained by its application will be given in a subsequent paper.[7] In this paper there will only be given a statement of the principles involved.

Since the network is the electrical analogue of the field problem, it is expected that its Q and input resistance should be related to those of a cavity resonator it represents. Normally, the network will correspond to a cavity with a nonuniform lossy dielectric, the series resistance of the inductances corresponding to the dielectric loss To represent wall losses, it is necessary to shunt the terminal capacitances of the network with resistances which are large compared to the capacitor reactance and which increase linearly with radius.

In making measurements upon a network loaded with terminal resistors to represent wall losses, it is necessary to separate the effects of the power losses in the termina resistors and in the coils themselves. For Q measurements, the Q of the network must be measured with and without the terminal resistors, thus determining the Q resulting from the terminal resistors alone. This value will be proportional to the cavity-resonator Q. The appropriate proportionality factor may be measured or deduced theoretically.

In making shunt-resistance measurements, use can be made of the fact that for TEM and TM_0 modes, the network quantities are the inverse of the field quantities thus the electric field is represented by a network current. Accordingly, it is expected that the equivalent circuit of the network, as measured at the cavity gap should be the inverse of the cavity-resonator equivalent circuit. Since the equivalent circuit of a cavity is given closely by a parallel combination of shunt resistance lossless inductance, and capacitance, it is expected that the equivalent circuit of the network will be a series combination of resistance, inductance, and capacitance This series impedance is measured by opening the network at the point where the radial gap currents flow into the axis circuit. It is further expected that the conductance of the series network equivalent will be proportional to the shunt resistance of the cavity. Here again, it is necessary to separate components of shunt resistance due to wall losses and dielectric losses. This is done by measuring the network series resistance with

[7] F. W. Schott and K. R. Spangenberg, "A network analogue approach to the study of cavity resonator losses," presented, 1948 West Coast IRE Convention, Los Angeles, Calif., October 1, 1948.

d without the terminal resistors. The conductance of e difference of the two series resistances thus measured proportional to the shunt resistance of the cavity.

Use of the Rectangular Network

1. Determination of Resonant Frequencies of Cavities. ie method used here involves excitation and variation frequency until electrical resonance is observed by the me method used on the cylindrical-co-ordinate net-ork. Attention must again be given to proper bound-es and end corrections. As with the other network, the ective field edge is half a section beyond the capacitor which the network is opened with open-circuit bound-es, and at the short with short-circuit boundaries.

2. Tuning Curves. These are obtained by the same thod as is used with the cylindrical-co-ordinate net-ork.

3. Field Plots. These can usually be obtained di-ctly from network voltage measurements. In the case the modes referred to here as *TM* modes, network ltage to ground is directly proportional to electric ld strength. A plot of contours of constant voltage to ound constitutes a magnetic flux plot. With *TE* odes, voltage to ground gives strength of magnetic ld, and contours of constant voltage constitute an ctric-field plot.

4. Measurement of Shunt Resistance and Q. Shunt re-tance and Q can be measured by the same general thod indicated for cylindrical cavities. In the case of veguide cavities operating on *TM* modes, however, e coil losses correspond to top and bottom surface ses. As a result, terminal resistors must be propor-ned relative to the coil resistance to represent prop-y side-wall losses. Also, care must be taken in repre-ting losses associated with gap loading, which, be-se of high current densities near the gap, will often as large as all other losses combined.

5. Measurement of Ridge Waveguide Characteristics. ork on ridge waveguides has indicated a number of irable characteristics.[8] These include low cutoff fre-encies and increased separation of modes. Other char-eristics, such as increased attenuation and lower ver-transmission capabilities, are less favorable. The aracteristics of ridge waveguides can readily be deter-ned with the rectangular network. The network, be-restricted to two-dimensional problems, can only e the field configuration at cutoff; however, with h the field configuration and cutoff frequency known, of the characteristics of the waveguide at any fre-ency can be determined.[9]

The principal characteristics of interest in ridge wave-des are the cutoff frequencies of the various modes l the impedance of the fundamental mode. Cutoff frequencies are readily determined by simply setting up the waveguide in cross section with open boundaries, according to the *TE* analogy previously given. The network is then excited through a high series resistance to give the effect of loose coupling to a variable-frequency oscillator. The cutoff frequencies of the various *TE* modes of transmission are scaled by equation (12) in Part 1 from the frequencies at which the network resonates.

The characteristic impedance of a waveguide may be defined in terms of (a) equivalent current and voltage; (b) transmitted power and equivalent voltage; or (c) transmitted power and equivalent current.[10] The current-voltage definition is always the geometric mean of the other two. The spread of values is of the order of 2 to 1 in ordinary waveguide, and is expected to be relatively small in ridge waveguide because this approaches a strip transmission line for which the three above impedances are identical. The current-voltage definition is most useful in considering matching problems with two-conductor lines, and agrees most closely with the values measured through transducers from concentric line to waveguide. All of the impedance forms contain a factor which depends solely upon the geometry of the waveguide multiplied by a frequency-dependent term of the form $(1-\eta^2)^{1/2}$ where η is the ratio of the cutoff frequency of the mode considered to the actual frequency. This factor is infinite at the cutoff frequency and drops to unity at infinite frequency.

The current-voltage definition of impedance is

$$Z_{I,V} = \frac{\int_b^t E_{y\max} dx}{\int_L^R H_x dx} \tag{21}$$

where the integral limits b, t, L, and R stand for bottom, top, left, and right, respectively, as shown in Fig. 8. All of the above quantities are simply related to the x

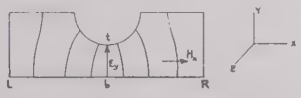

Fig. 8—Ridge-waveguide notation.

component of magnetic field of which the network analogue is voltage to ground. Specifically,

$$E_y = \frac{j\beta Z}{K^2} \frac{\partial H_z}{\partial x} \tag{22}$$

$$H_x = \frac{-j\beta}{K^2} \frac{\partial H_z}{\partial x} \tag{23}$$

where β is the phase factor, k is $2\pi/\lambda$, and $Z = (1-\eta^2)^{-1/2}$. With these substitutions, and replacing H_z by V, net-

[8] S. B. Cohn and P. Richards, "Very High Frequency Techues," McGraw-Hill Book Co., Inc., New York, N. Y., 1947; II, pp. 678–684.

[9] J. C. Slater, "Microwave Transmission," McGraw-Hill Book Inc., New York, N. Y., 1942; p. 152.

[10] S. A. Schelkunoff, "Electromagnetic Theory," D. Van Nostrand Co., Inc., New York, N. Y., 1943; pp. 319–324.

work voltage to ground, the impedance expression becomes

$$Z_{VI} = 337 \frac{(1-\eta^2)^{-1/2} \int_b^t \frac{\partial V}{\partial x} dy}{V_R - V_L} \text{ ohms.} \qquad (24)$$

From this it is seen that the impedance of the waveguide is given by 377 times the frequency factor, times the average transverse voltage gradient at the waveguide center, multiplied by the vertical dimension of the waveguide; all divided by twice the maximum voltage which appears on the edge of the network. Impedance values according to the other definitions are similarly obtained.

C. Solution of Radiation Problems

The networks may be used to solve antenna problems, give radiation patterns, surface currents, antenna impedances, junction transformations, etc. To simulate radiation problems, it is necessary to simulate the antenna contour and then terminate the edges of the network in such a way as to eliminate reflection of waves from it. Experiments with network terminations have shown that standing waves of the order of 1.10 or less are readily obtainable. This makes possible the determination of radiation patterns. Antenna impedances can be measured directly. Reactance transformation of junctions can be tested by simulating the surrounding of the antenna with reflecting spheres of different radii. For extensive work with antenna patterns, a special network containing a centrally located fine section with larger dimensions over-all in terms of wavelengths, and possibly having a polar co-ordinate system rather than a rectangular co-ordinate system, would be advantageous.

D. Accuracy and Speed of Network Determinations

The networks described here are not high-precision devices, but do give sufficient accuracy for engineering applications. Resonant frequencies may be determined to within 1 per cent accuracy for simple geometries or 2 per cent accuracy for complex geometries. Shunt resistance and Q determinations are accurate to within 5 per cent for complex geometries, but may be relied upon to be within 2 per cent accurate for simple geometries. This accuracy decreases somewhat as the electrical length of a section exceeds 25°.

A complete field plot of a resonator may be obtained in less than an hour. Resonant frequencies may be determined within a matter of minutes. Once equipment has been set up for shunt-resistance measurements, such determinations may also be made in a matter of minutes. In general, determinations are much more rapid than the equivalent calculation or model measurement.

VI. Experimental Verification of Performance

A. TEM Modes

A large number of resonant structures have been checked against the network to calibrate its operation. Checks on resonant frequency, field configuration, shunt resistance, and Q have been made. A few of these results are reported here to indicate the nature of agreement between the theory and the network results.

A basic reference test is the resonant concentric line operating on a *TEM* mode with its length being an integral number of half waves. Such a configuration when set up, gives resonant frequencies which check within 1 to 2 per cent. The agreement on the field variation with line length is correspondingly good. Shown in Fig. 9 are plots of measured and theoretical

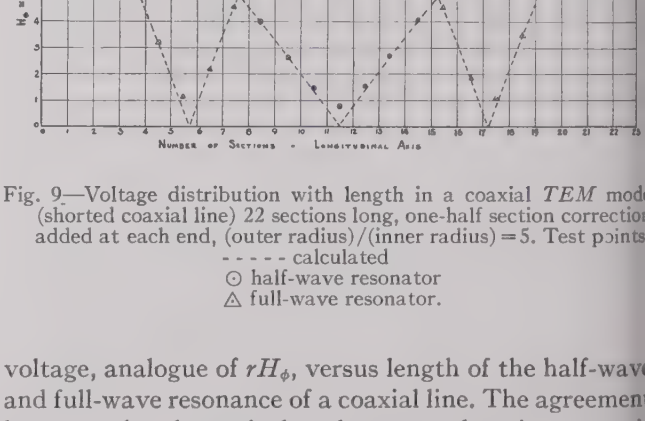

Fig. 9—Voltage distribution with length in a coaxial *TEM* mode (shorted coaxial line) 22 sections long, one-half section correction added at each end, (outer radius)/(inner radius) = 5. Test points
- - - - - calculated
⊙ half-wave resonator
△ full-wave resonator.

voltage, analogue of rH_ϕ, versus length of the half-wave and full-wave resonance of a coaxial line. The agreement between the theoretical and measured cosine wave is good except in the vicinity of the nodes, where the true reading is somewhat obscured by harmonics and noise. This particular mode is convenient for checking the network adjustment on the cylindrical network. Since magnetic field varies inversely with radius, network voltage which is proportional to rH_ϕ, will be constant with radius. It is very easy to check this condition with a vacuum-tube voltmeter, and any irregularities can usually be quickly traced. From resonant-frequencies measurements, the half-section correction for open-circuit boundaries can be verified.

B. Cylindrical TM_{0m0} Modes

The cylindrical-resonator TM_{0m0} modes are formed by radial waves which have no variations in intensity in the axial direction. The corresponding resonant frequencies are inversely proportional to radius. Tests with a large number of radii showed that measured network frequencies agreed with theoretical values within about 1 per cent when the radial-field edge correction was con-

ered to be half a network section. For these modes, radial electric-field variation is expected to be a zero-er Bessel function of the first kind. The magnetic d is expected to vary as a first-order Bessel function he first kind. Figs. 10 and 11 show that these expec-ons were fulfilled quite closely.

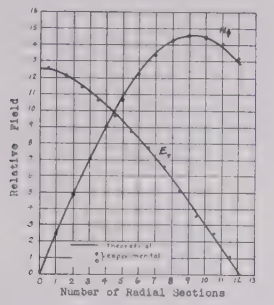

Fig. 10—Radial field variation in the TM_{010} mode.

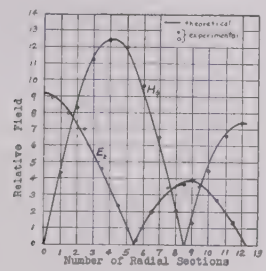

Fig. 11—Radial field variation in the TM_{020} mode.

TM_{011} Mode

The cases checked above have shown that the cylin-al-co-ordinate network behaves properly for purely ial or longitudinal waves. The question naturally es as to whether it will properly represent a com-ation of these. The simplest mode which exhibits h radial and axial field variations is the TM_{011} mode. s mode is found with the same boundary conditions he TM_{010} mode, it being necessary only to raise the uency of the excitation voltage to the proper value. excitation frequency in a typical experimental test 136.8 kc, as against a theoretical value of 135.3 kc.

Net-Point Computations

n order to check the performance of the network on nator structures presenting more complex geome-s than the simple forms shown above, it is necessary have recourse to rather involved theoretical ap-ximations or resort to net-point computations. Net-nt computations were preferred in checking network lts because they are adaptable to arbitrary geo-rical configurations, and also because the network e the resonant frequency with sufficient accuracy t the computations were greatly simplified.

Basically, the net-point computation consists of find-a difference equation which corresponds to the dif-ntial equation pertinent to the problem; assuming ues of field and then applying the difference equation to obtain successive improvements on the originally assumed values of field.[11–12]

In making net-point computations of field in resonators with rotational symmetry, it is convenient to use the parameter rH_ϕ, which is the analogue of network voltage to ground, V. This is convenient because, as has been shown, contours of constant rH_ϕ or V constitute an electric-field plot. Likewise, values of rH_ϕ or V divided by radius give magnetic field. The boundary conditions on rH_ϕ or V are also simple in that the normal gradient of rH_ϕ is zero at any conducting surface. The differential equation for V in two-dimensional, cylindrical co-ordinates with rotational symmetry has the form

$$\frac{\partial^2 V}{\partial z^2} + \frac{\partial^2 V}{\partial r^2} - \frac{1}{r}\frac{\partial V}{\partial r} + k^2 V = 0 \qquad (25)$$

where k is 2π divided by wavelength. Note that this equation differs properly from the scalar wave equation in that the sign of the first derivative term in r is negative. The corresponding net-point or difference equation in terms of the network notation of Fig. 12 is

$$V_0 = \frac{V_1 + V_2 + V_3 + V_4 - \dfrac{(V_4 - V_3)}{2I}}{4 - h^2 k^2} \qquad (26)$$

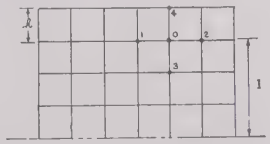

Fig. 12—Notation for net-point computations in cylindrical co-ordinates.

where h is the net-point spacing in wavelengths and I is the integral number of lattice sections from the axis. This equation relates the voltage in the center of a square to the voltage at four symmetrically disposed points arranged as shown in Fig. 12. If values of voltage are assumed at all points in a field or are taken from actual network measurements, then application of (26) will smooth out the values and give values which are more correct than those originally given.

Shown in Table VI are values of network voltage as taken directly from the network (above), compared with voltage values refined by a successive application of (25). It is seen that the errors in the field as presented in the network are relatively slight.

For cases in which the resonant frequency is not known, it is necessary to reiterate between values of field and frequency. However, the dependence of the field values upon the frequency is low. Hence if the resonant frequency is known within a few per cent, it is

[11] G. H. Shortley and R. Weller, "Numerical solution of Laplace's equation," *Jour. Appl. Phys.*, vol. 9, pp. 334–348; May, 1948.

[12] H. Motz, "Calculation of the electromagnetic field, frequency, and circuit parameters of high-frequency resonator cavities," *Jour. IEE* (London), vol. 93, part III, pp. 335–338; September, 1946.

possible to proceed to refine field values without correcting frequency.

In some applications, it is desired to use a different form of the difference equation given above. Where extensive computations are to be made, it is convenient to have a form where the center potential of a square is given by the sum of the four surrounding values divided by a constant dependent upon the radius. Such a form results if there is used the variable

$$U = V/\sqrt{r}. \tag{27}$$

Another useful transformation is

$$W = V/r^2. \tag{28}$$

The paramenter W has the advantage that it will tend to be constant in a resonator gap on the axis. Since V is proportional to rH_ϕ, and H_ϕ tends to increase linearly with radius in a gap, V/r^2 will tend to be constant.

It should be emphasized that the tests reported here and many others indicate a good agreement between the analyzer results and theory in cases where the results may be simply predicted and also in cases where the theoretical results are not simply predicted. As a result, it is felt that the analyzer is a useful tool which can be trusted in cases where the boundaries are too complicated to admit of only experimental or numerical determination.

VII. Conclusions

Networks have been constructed which give the solu tion of the wave equation in two-dimensional rectangu lar and cylindrical co-ordinates. Resonant frequencie of cavities, field plots, equivalent reactance of obstacle shunt resistance, and Q are obtainable with little effor The devices are simple to operate, and results can b obtained with considerable speed. The networks contai no vacuum tubes or moving parts and are simple an inexpensive to build. Accuracy, though not at the pr cision level, is sufficient for most engineering applic tions.

Two-dimensional networks are believed to be ver practical. It is not thought that three-dimensional ne works are practical.

The educational value of the networks is considera ble. They possess a form of "intelligence," in that the often reveal the presence of unsuspected resonanc modes. In addition, their use reveals a large number o important properties of fields which are little know They can be operated by people not highly skilled i field theory.

VIII. Acknowledgment

The assistance of E. A. Goddard, P. Tissot, J. F Honey, R. Tanner, and M. B. Adams with variou phases of the work is gratefully acknowledged.

TABLE VI

	A	B	C	D	E	F	G	H	I	J	K	
12												
	120 121	160 100	235 235	345 346	475 475	625 625	780 780	930 930	1090 1090	1190 1190		1250 1250
11												
	72.0 68.3	105 103.1	175 174.5	280 280	400 400	540 540	630 630	840 840	980 980	1070 1060		1120 1120
10												
	22.5 23.2	10.0 2.2	57.0 63.8	157 155	265 266	390 390	515 515	640 640	755 755	820 820		890 890
9												
	140 141	113 119.1	62.0 75.0	13.0 4.13	105 37.3	210 202.5	320 314	420 428	515 523	590 606		690 645
8												
	260 255	240 236.5	220 200	148 153.8	80.0 84.0	11.5 4.3	103 96.4	138 209	280 275	340 332		370 382
7												
	335 338	330 323	310 310	280 283.5	250 242	187 187.5	117 114.3	44.0 38.5	23.5 27.4	80.0 77.7		105 103.
6												
	365 369	370 371	370 371	370 363	362 330	330 330	282 285	230 231	180 178.5	140 138.4		120 116.
5												
	360 353	365 363	385 375	420 422	440 435	415 428	400 399	360 359	315 318	280 285		265 268
4												
	285 287	295 300	335*	425 429	470 470	470 470	450 453	415 421	380 385	355 355		340 339
3												
	187 189.9	191 195.5	202 205									
2												
	92.0 94.1	94.0 95.7	98.0 98.2									
1												
	24.0 25.4	25.0 25.6	25.0 25.9									
0												
	0*	0*	0*									

* This value kept constant.

Design of Optimum Transient Response Amplifiers*

PIERRE R. AIGRAIN†, STUDENT MEMBER, IRE, AND EVERARD M. WILLIAMS‡, SENIOR MEMBER, IRE

Summary—**This paper describes a method based entirely upon nsient considerations for the design of amplifiers with optimum nsient response in the "least squared" sense. This method is licable to video amplifiers, symmetrical band-pass amplifiers, l dissymmetrical band-pass amplifiers used with low-level modu-on, and has provisions for taking into account noise acceptance l adjacent channel rejection. Examples are given of the results ained with this method.**

Introduction

ALTHOUGH ALL communication amplifiers are used in conjunction with signals of an essentially transient character, amplifier circuit design and nthesis has for the most part concentrated on the oblem of obtaining predetermined steady-state char-teristics. Methods of solving this steady-state prob-n have been brought to a high degree of perfection by uer,[1] Guillemin,[2] and Lee,[3] amongst others.

Recent developments, however, first of television, er of radar, and finally of pulse communication sys-ns, showed that there are important cases where the rformance of a system cannot be simply related to its ady-state characteristics, but rather is best expressed its transient response.

Existing design methods for transient amplifiers are rived from steady-state theories, particularly those ating to filters. Butterworth,[4] Landon,[5] Wallman,[6] dford,[7] and Kallman, Spencer, and Singer[8] all studied plifier transient response as related to steady-state aracteristics. The shortcomings of their procedures e well understood, however, and have been pointed t by Kallmann.[8]

This paper describes a design method, based entirely on transient analysis, which leads to amplifiers with an optimum transient response. This method is derived from operational analysis using Laplace transforms.

The procedure by which the transient response of an amplifier to some particular signal is calculated with Laplace transform technique involves the determination of the "operational transfer impedance," $AG(s)$, of the circuit under consideration. If a signal of Laplace transform $F_1(s)$ is applied to the amplifier, the output will have the transform

$$AF_2(s) = AG(s)F_1(s). \tag{1}$$

(The amplifier gain A, at low frequencies, is extracted as a coefficient for convenience.) Considerable difficulty is generally encountered when it is desired to calculate the actual amplifier response $f_2(t)$, the inverse transform of $F_2(s)$ in (1). This process requires the extraction of the poles of $G_2(s)$, an algebraic process only possible in general with known numerical values and of great difficulty when several poles are involved. It is apparent first that a simpler method is necessary for calculating the transient response. Such a method is described in part I.

I. The Laguerre Series

The problem is that of determining $f(t)$ from a specified $F_2(s)$. This will be simplified first by using a function $F_3(t)$, defined as

$$f_3(t) = f_2(t) - C$$
$$C = f_2(\infty) = \lim_{s \to 0} sF_2(s) \tag{2}$$

or a function differing from $f_2(t)$ by the steady-state response $f_2(\infty) \cdot f_3(t)$ can be calculated as

$$f_3(t) = \sum_{n=0}^{\infty} \alpha_n B_n(t), \tag{3}$$

in which the terms α_n are the coefficients of an expansion in a Taylor series of

$$\frac{2}{1-x} F_3\left(\frac{1+x}{1-x}\right) = \sum_{n=0}^{\infty} \alpha_n x^n$$

and the terms $B_n(t)$ are "Laguerre functions"[9] defined by

$$B_n(t) = e^{-t} \sum_{p=0}^{n} \frac{(-2)^p}{p!} \binom{n}{p} t^p. \tag{4}$$

The series (2) is convergent for all values of t. Since

$$\frac{2}{1-x} F_3\left(\frac{1+x}{1-x}\right)$$

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‡ Carnegie Institute of Technology, Pittsburgh, Pa.

[1] W. Cauer, "Siebschaltungen," published by V.D.I., Verlag M.B.H., Berlin; 1931.

[2] E. A. Guillemin, "Communication Networks," John Wiley and s, Inc., New York, N. Y., 1931.

[3] Y. W. Lee, "Synthesis of electric networks by means of the rier transform of Laguerre's functions," *Jour. Math. Phys.*, vol. pp. 83–113; 1932.

[4] S. Butterworth, "On the theory of filter-amplifiers," *Exp. Wire-and Wireless Eng.*, vol. 7, pp. 536–541; October, 1940.

[5] V. D. Landon, "Cascade amplifiers with maximal flatness," *A Rev.*, vol. 5, pp. 347–362; January, 1941.

[6] H. Wallman, "Stagger Tuned I.F. Amplifiers," MIT Radiation . Report 524; February, 1944.

[7] A. V. Bedford and G. L. Fredendall, "Transient response of lti-stage video frequency amplifiers," PROC. I.R.E., vol. 27, pp. –289; April, 1939.

[8] H. E. Kallman, R. E. Spencer, and C. P. Singer, "Transient ponse," PROC. I.R.E., vol. 34, pp. 138–144; March, 1946.

[9] The Laguerre functions are related to the Laguerre polynomials, for which see "Orthogonal Polynomials", Szegö, Gabor, Colloquium Publications, American Mathematical Society, vol. 23; 1939.

is a rational fraction in x, α_n can be obtained by division of two polynomials. Furthermore, the $B_n(t)$ form a close orthogonal set from $t=0$ to $t=\infty$, that is

$$\int_0^\infty B_n B_m dt = \begin{cases} 0 & \text{if } n \neq m \\ \frac{1}{2} & \text{if } n = m \end{cases} \tag{5}$$

and every function $L(t)$ such that

$$\int_0^\infty L(t) B_n(t) dt = 0 \quad \text{for all } n \tag{6}$$

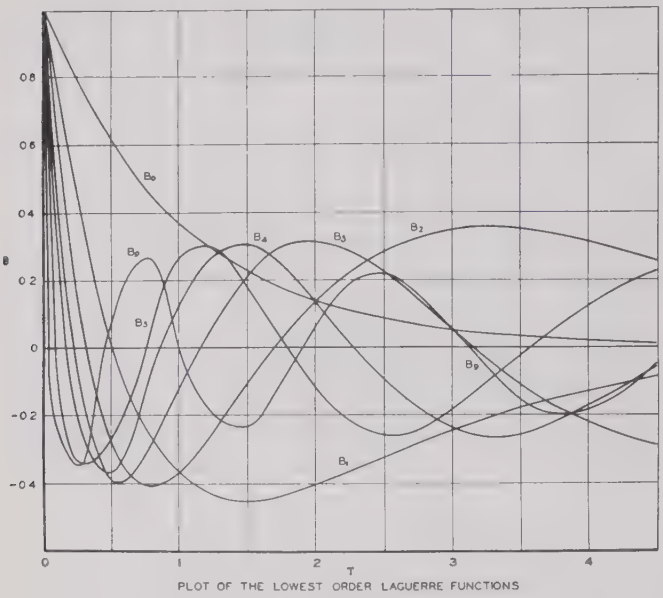

Fig. 1—Plot of the first nine Laguerre functions.

is identically zero almost everywhere. The first ten $B_n(t)$ have been computed and are given in Table I. Their form is shown in Fig. 1. Calculation of transient response is often simplified by the use of these $B_n(t)$ and (2) and (3); the principal value of these functions, however, is that they result in a simple means for determining the effect of individual parameter variations on the over-all transient response. This subject will be treated in part III. Part II deals with the definition of transient distortion.

II. Definition of Transient Distortion

The distorting or disturbing effects of principal inte est in communication systems are non-linear distortio random noise, adjacent channel interference, and tran ent distortion. When defining the first three it is cu tomary to relate the actual response $f_q(t)$ to a specifi signal to the output $f_p(t)$ of a "distortionless" syste The actual distortion is defined as the time average of

$$[f_p(t) - f_q(t)]^2$$

or, if the input is of finite duration, as

$$\int_{-\infty}^{\infty} [f_p(t) - f_q(t)]^2 dt.$$

This definition has given satisfactory results when a plied to most types of distortion and it seems logical extend it to transient distortion. It has, in fact, prev ously been so used.[10] Following this concept, transie distortion will be defined as follows: Assume a sign $f_1(t)$ to be fed into the amplifier at time $t=0$. The outp will be $A f_2(t)$. An ideal amplifier would yield an outp $A f_1(t-T_0)$ in which T_0 represents a delay time. Th "transient distortion" will be defined as

$$I = \int_0^\infty [f_2(t) - f_1(t - T_0)]^2 dt. \tag{7}$$

This integral is convergent for all $f_1(t)$ with a finite e ergy and for all bounded input functions[11] which rea a final value in a finite time (e.g., a step function), pr vided the amplifier has no low frequency distortion. Th low-frequency distortion problem in video amplifie however, is fairly distinct from that of high-frequen distortion, so that it is possible to define an "equivale low-pass amplifier" without low-frequency distortio and work with it exclusively.

TABLE I

Values of the First Nine Laguerre Functions

t	$B_0(t)$	$B_1(t)$	$B_2(t)$	$B_3(t)$	$B_4(t)$	$B_5(t)$	$B_6(t)$	$B_7(t)$	$B_8(t)$	$B_9(t)$	$B_{10}(t)$
0.25	0.7788	0.3894	0.0974	−0.1136	−0.2576	−0.3470	−0.3926	−0.4026	−0.3881	−0.3527	−0.303
0.50	0.6065	0	−0.3033	−0.4043	−0.3791	−0.2830	−0.1558	−0.0243	0.0935	0.1883	0.255
0.75	0.4724	−0.2362	−0.4133	−0.3248	−0.1366	0.0550	0.2008	0.2828	0.3016	0.2679	0.197
1.00	0.3679	−0.3679	−0.3679	−0.1226	0.1226	0.2698	0.3025	0.2441	0.1320	−0.0391	−0.153
1.25	0.2865	−0.4298	−0.2507	0.0776	0.2753	0.2958	0.1897	0.0309	−0.1176	−0.2171	−0.252
1.50	0.2231	−0.4463	−0.1117	0.2331	0.3068	0.1897	−0.0028	−0.1666	−0.2473	−0.2368	−0.156
1.75	0.1738	−0.4344	0.0217	0.3005	0.2467	0.0310	−0.1668	−0.2507	−0.2160	−0.1013	0.030
2.00	0.1353	−0.4060	0.1353	0.3158	0.1353	−0.1174	−0.2511	−0.2206	−0.0847	0.0737	0.180
2.25	0.1054	−0.3689	0.2240	0.2833	0.0091	−0.2185	−0.2443	−0.1093	0.0701	0.1991	0.228
2.50	0.0821	−0.3283	0.2873	0.2189	−0.1060	−0.2599	−0.1715	0.0267	0.1836	0.2206	0.144
2.75	0.0639	−0.2877	0.3276	0.1372	−0.1943	−0.2457	−0.0634	0.1427	0.2250	0.1578	0.014
3.00	0.0498	−0.2490	0.3485	0.0498	−0.2490	−0.1892	0.0498	0.2120	0.1952	0.0498	−0.110
3.25	0.0388	−0.2132	0.3538	−0.0347	−0.2693	−0.1060	0.1467	0.2295	0.1194	−0.0682	−0.187
3.50	0.0302	−0.1812	0.3473	−0.1107	−0.2605	−0.0156	0.2067	0.1905	0.0097	−0.1586	−0.199
3.75	0.0235	−0.1529	0.3323	−0.1750	−0.2273	0.0718	0.2313	0.1202	−0.0897	−0.2015	−0.151
4.00	0.0183	−0.1282	0.3114	−0.2259	−0.1771	0.1295	0.2202	0.0327	−0.1640	−0.1932	−0.065

[10] W. W. Hansen, "Transient response of wide-band amplifiers *Proc. Nat. Elect. Conference*, vol. 1, pp. 544–553; October, 1944.

[11] If the input is a delta (δ) impulse, the definition of transient d tortion must be modified slightly to yield a convergent result. suitable definition is

$$I = \int_0^\infty [f_2^2(t) - 2\delta(t - T_0) f_2(t)] dt = \int_0^\infty f_2(t) dt - 2 f_2(T_0).$$

In this study, in accordance with common practice in eo amplifiers, $f_1(t)$, will be taken as a step function de- ed by

$$f_1(t) = 0 \quad \text{for} \quad t < 0$$
$$f_1(t) = 1 \quad \text{for} \quad t > 0.$$

e method can, if desired, readily be extended to other nals. The delay T_0 in (7) must be chosen to minimize For the step function

$$f_2(T_0) = \tfrac{1}{2}.$$

III. Calculation and Minimization of I

The transient distortion integral I can very easily be luated when the Laguerre series expansion of $f_3(t)$ is d. In fact, if we write

$$f_1(t - T_0) = \sum_{n=0}^{\infty} \gamma_n B_n(t) + 1,$$

n, using $f_3(t)$, from (2),

$$I = \sum_{n=0}^{\infty} (\alpha_n - \gamma_n)^2. \tag{8}$$

is formula permits the evaluation of the integral I hout calculating $f_2(t)$ explicitly, providing values of T_0) are available. Table II gives these values for the e in which $f_1(t)$ is a step function.[12]

pendent parameters $\xi_1, \xi_2, \xi_3 \cdots \xi_n$. The integral I may be minimized when

$$\xi_1 = C_1, \xi_2 = C_2, \cdots, \xi_n = C_n,$$

if

$$\frac{\partial I}{\partial \xi_1}(C_1, C_2, \cdots, T_0) = 0$$
$$\frac{\partial I}{\partial \xi_2}(C_1, C_2, \cdots, T_0) = 0, \text{ etc.} \tag{9}$$

The series (3) and (4) are absolutely convergent, and (8) may be differentiated term by term. The optimizing equations corresponding to (9) are:

$$\sum_{n=0}^{\infty} (\alpha_n - \gamma_n) \frac{d\alpha_n}{d\xi_1} = 0. \tag{10}$$

Since the Laplace transform of the derivative of a function of time with respect to an independent parameter is the derivative of the Laplace transform of this function with respect to the parameter,

$$-\frac{\partial}{\partial \xi_1} \frac{2}{1+x} G\left(\frac{1+x}{1-x}\right) = \sum_{n=0}^{\infty} \frac{\partial \alpha_n}{\partial \xi_1} x^n. \tag{11}$$

This yields $\partial\alpha_n/\partial\xi_1$ just as (3) yields α_n. The system (11) solves the problem mathematically, but actually an approximate method is desirable for its solution. We as-

TABLE II

Values of $\gamma_p(T_0)$ for the Case of the Step Input Function

T_0	$\gamma_0(T_0)$	$\gamma_1(T_0)$	$\gamma_2(T_0)$	$\gamma_3(T_0)$	$\gamma_4(T_0)$	$\gamma_5(T_0)$	$\gamma_6(T_0)$	$\gamma_7(T_0)$	$\gamma_8(T_0)$	$\gamma_9(T_0)$	$\gamma_{10}(T_0)$
.25	0.442	0.336	0.248	0.174	0.114	0.065	0.026	−0.006	−0.023	−0.048	−0.051
.50	0.787	0.426	0.181	0.021	−0.072	−0.120	−0.134	−0.129	−0.107	−0.083	−0.051
.75	1.055	0.362	−0.008	−0.171	−0.203	−0.180	−0.112	−0.052	0.015	0.053	0.088
.00	1.264	0.207	−0.207	−0.283	−0.207	−0.087	0.024	0.091	0.133	0.198	0.137
.25	1.427	0.006	−0.364	−0.293	−0.103	0.062	0.151	0.167	0.130	0.069	0.003
.50	1.557	−0.215	−0.454	−0.215	0.048	0.186	0.199	0.129	0.053	−0.074	−0.097
.75	1.672	−0.476	−0.436	−0.121	0.229	0.201	0.195	−0.027	−0.052	−0.167	−0.096
.00	1.729	−0.647	−0.436	0.079	0.278	0.227	0.041	−0.102	−0.170	−0.147	−0.079
.25	1.789	−0.841	−0.345	0.227	0.322	0.133	−0.082	−0.188	−0.170	−0.088	−0.029
.50	1.836	−1.015	−0.216	0.353	0.297	0.011	−0.188	−0.209	−0.105	0.031	0.122
.75	1.872	−1.189	−0.022	0.403	0.260	−0.158	−0.207	−0.205	0.041	0.094	0.193
.00	1.900	−1.303	0.108	0.490	0.108	−0.228	−0.250	−0.074	0.108	0.183	0.172
.25	1.922	−1.418	0.284	0.493	−0.023	−0.303	−0.202	0.037	0.184	0.192	0.047
.50	1.940	−1.517	0.460	0.456	−0.157	−0.333	−0.111	0.144	0.218	0.119	−0.038
.75	1.953	−1.600	0.630	0.385	−0.280	−0.318	−0.001	0.223	0.197	0.047	−0.168
.00	1.963	−1.670	0.791	0.283	−0.381	−0.232	0.051	0.324	0.069	−0.011	−0.245

s necessary, however, to evaluate T_0 (to avoid an ex- cit solution) and a simplified method of calculation is en subsequently. Equation (8) not only permits the culation of I but also provides a means for choosing cuit parameters to render I a minimum. In general. the coefficients of $G(s)$ cannot be chosen arbitrarily. ne relations between them are introduced by ampli- gain requirements and by unavoidable minimum ues of interstage shunt capacitances. It is convenient express the coefficients of $G(s)$ as functions of inde-

sume that approximate values $C_{10}, C_{20}, \cdots, C_{n0}$ can be found for $C_1, C_2, \cdots, C_n$ so that $\Delta_1, \Delta_2, \cdots$, defined by

$$\Delta_1 = C_1 - C_{10},$$
$$\Delta_2 = C_2 - C_{20},$$

are small quantities.

The equations (10) can be expanded in a series of powers of $\partial\alpha_n/\partial\xi_1$ and, neglecting all of order above the first, this series is

$$\sum_{n=0}^{N} (\alpha_n - \gamma_n) \frac{\partial \alpha_n}{\partial \xi_1}$$

[12] If $f_1(t)$ is a delta impulse, $\gamma_n = B_n(T_0)$.

$$+\Delta_1\left\{\sum_{n=0}^{N}\left(\frac{\partial\alpha_n}{\partial\xi_1}\right)^2+(\alpha_n-\gamma_n)\frac{\partial^2\alpha_n}{\partial\xi_1^2}\right\}$$

$$+\Delta_2\left\{\sum_{n=0}^{N}\frac{\partial\alpha_n}{\partial\xi_1}\cdot\frac{\partial\alpha_n}{\partial\xi_2}+(\alpha_n-\gamma_n)^2\frac{\partial^2\alpha_n}{\partial\xi_1\partial\xi_2}\right\}=0, \quad (12)$$

and other similar equations. This system (12) is linear and readily solved. The number N in this system corresponds to the last series terms considered to be of importance. Since numerous approximations are involved in (12), the values of $\xi_1, \xi_2, \cdots$, obtained will differ from $C_1, C_2, \cdots$. However, the amount by which ξ_1, differs from C_1, etc., is not so significant as the amount by which the transient distortion I of the calculated circuit differs from that of the best possible circuit. The approximations can be shown to introduce the following errors in I:

1. An error ϵ_1, due to the neglect of terms of order above the Nth:

$$\epsilon_1 < 2\left|\sum_{n=0}^{\infty}2\alpha_n\gamma_n-\alpha_n^2\right|.$$

2. An error ϵ_2, due to taking for T_0 an approximate value,

$$\epsilon_2 < 0.01I \quad \text{if} \quad 0.47 < f_2(T_0) < 0.53$$
$$< 0.001I \quad \text{if} \quad 0.49 < f_2(T_0) < 0.51.$$

3. An error ϵ_3, due to the linear approximation to equation (10). This error cannot be ascertained easily, but is of the third order in $\Delta_1, \Delta_2, \cdots$, so that it is very small if $\Delta_1, \Delta_2, \cdots$ are small.

There remains the problem of determining $C_{10}, C_{20}, \cdots$. Any approximate design methods may be utilized. Marcuwitz[13] has proposed a method in which the first terms in the Laplace transform transfer impedance $G(s)$ in powers of s are equated to those of the series expansion of e^{-sT_0}. If all the terms could be equated a perfect transient response could result and further, if circuits of increasing complexity are considered, so that more and more terms can be equated, results of uniformly increasing quality are obtained. Although this method does not necessarily lead to an approximately optimum circuit in any given case with a finite number of terms, it does lead consistently to usable designs as a starting point for our optimization procedures. The Marcuwitz method may be replaced by any other approximate design data.

As an illustrative application consider the circuit of Fig. 2. This interstage coupling network has been studied by Kallmann, Spencer, and Singer[8] for equal input and output capacitances. We shall determine the optimum values of the parameters subject to choice with the sum of input and output capacitances fixed, but with an additional choice as to the division of this total capacitance between input and output circuits; that is, with

[13] N. Marcuwitz, "Distortionless Correction of Video Networks," M.E.E. Thesis, Polytechnic Institute of Brooklyn, June, 1941.

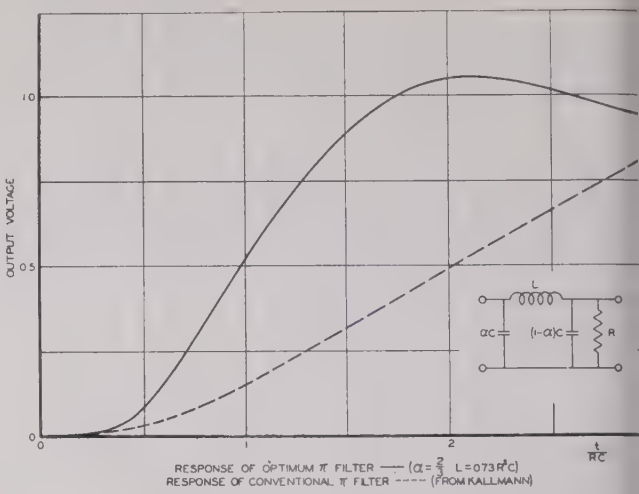

Fig. 2—Pi-filter interstage coupling network. Responses of the conventional type and the optimum type are compared.

fixed values of C and R (R from the gain selected) an variable L and α_c. The time unit is chosen as $t/(RC)$ s that a general solution is obtained. The notation $\xi_1=\alpha_c LC$ and $\xi_2=\alpha_c(1-\alpha_c)$ will be used.

The Laplace transform of this circuit is

$$G_{(s)}=\frac{1}{p(s)}=\frac{1}{1+s+\xi_1 s^2+\xi_2 s^3},$$

and the transform of the response to a step function is

$$G_1(s)=\frac{1}{s}G(s)=\frac{1}{sp(s)}.$$

The derivatives of $G_1(s)$ with respect to ξ_1 and ξ_2 are

$$\frac{\partial G_1}{\partial\xi_1}=\frac{s}{p^2(s)} \qquad \frac{\partial G_1}{\partial\xi_2}=\frac{s^2}{p^2(s)}$$

$$\frac{\partial^2 G_1}{\partial\xi_1^2}=\frac{-2s^3}{p^3(s)} \qquad \frac{\partial^2 G_1}{\partial\xi_1\partial\xi_2}=\frac{-2s^4}{p^3(s)}.$$

As the starting point approximation, use the circui given by Kallmann, Spencer, and Singer[8] in whic $\xi_1=0.45$ and $\xi_2=0.10$. Substituting $S=(1+x)/(1-x)$ i the above expressions, with these values of ξ_1 and ξ_2

$$\frac{2}{1-x}G_1\left(\frac{1+x}{1-x}\right)=\frac{3.1-3.6x+1.5x^2}{2.55-3.25x+1.85x^2-0.35x^3}$$

$$\frac{2}{1-x}\frac{\partial G_1}{\partial\xi_1}=(1-x)M(x),$$

$$\frac{2}{1-x}\frac{\partial G_1}{\partial\xi_2}=(1+x)M(x)$$

$$\frac{2}{1-x}\frac{\partial^2 G_1}{\partial\xi_1^2}=(1-x)^2N(x),$$

$$\frac{2}{1-x}\frac{\partial^2 G_1}{\partial\xi_1\partial\xi_2}=(1-x^2)N(x)$$

$$\frac{2}{1-x}\frac{\partial^2 G_1}{\partial\xi_2}=(1+x^2)N(x),$$

which

$$\sum_0^\infty m_p x^p = M(x) = \frac{1 - 2x + 2x^3 - x^4}{3.25(1 - 2.55x + 3.07x^2 - 2.12x^3 + 0.87x^4 - 0.2x^5 + 0.02x^6)}$$

$$\sum_0^\infty n_p x^p = N(x) = \frac{1 - 3x^2 + 3x^4 - x^6}{4.14(1 - 3.82x + 7.04x^2 + 6.15x^4 - 3.27x^5 + 1.2x^6 - 0.29x^7 + 0.04x^8)}.$$

We can now arrange in Table III the coefficients of expansion of the above expressions in powers of x. course

$$\frac{\partial \alpha_p}{\partial \xi_1} = m_p - m_{p-1}, \qquad \frac{\partial \alpha_p}{\partial \xi_2} = m_p + m_{p+1},$$

$$\frac{\partial^2 \alpha_p}{\partial \xi_1^2} = n_p - 2n_{p-1} + n_{p-2}, \qquad \frac{\partial^2 \alpha_p}{\partial \xi_1 \partial \xi_2} = n_p - n_{p-2},$$

$$\frac{\partial^2 \alpha_p}{\partial \xi_2^2} = n_p + 2n_{p-1} + n_{p+2}.$$

IV. Extensions of the Method to Band-Pass Amplifiers and to the Treatment of Noise and Adjacent-Channel Rejection

The design method described so far applies only to video amplifiers. It can be extended to symmetrical band-pass amplifiers, or asymmetrical amplifiers with low-level modulation, however, by making use of the band-pass, low-pass analogy.[15,16] The design method here involves finding the low-pass equivalent of the band-pass amplifier under consideration, obtaining the optimum design for the video equivalent and reconverting to the band-pass equivalent.

TABLE III

p	γ_p	α_p	$\gamma_p - \alpha_p$	$3.25m_p$	$\frac{\partial \alpha_p}{\partial \xi_1}$	$4.19n_p$	$\frac{\partial \alpha_p}{\partial \xi_2}$	$\frac{\partial^2 \alpha_p}{\partial \xi_1 \partial_2}$	$\frac{\partial^2 \alpha_p}{\partial \xi_1^2}$	$\frac{\partial^2 \alpha_p}{\partial \xi_2^2}$
0	1.264	1.216	0.048	1.	0.308	−1.	0.308	−0.242	−0.242	−0.242
1	0.207	0.137	0.070	0.548	−0.139	− 3.82	0.475	−0.923	−0.437	−1.397
2	−0.207	−0.197	0.010	−1.678	−0.638	− 4.55	−0.347	−0.857	0.502	−3.166
3	−0.283	−0.184	−0.099	−1.839	−0.049	1.57	−1.076	1.302	1.644	−2.741
4	−0.207	−0.073	−0.134	−0.240	0.491	10.85	−0.638	3.72	0.758	2.014
5	−0.087	0.015	−0.102	1.202	0.442	14.39	0.295	3.12	−1.354	8.88
6	0.024	0.047	−0.023	1.458	0.079	9.12	0.816	−0.415	−2.162	11.82
7	−0.031	0.039	0.052	0.776	−0.209	− 1.26	0.685	−0.378	−1.200	7.30
8	0.133	0.018	0.115	−0.079	−0.262	−11.21	0.214	−4.88	0.103	−1.11
9	0.198	0.001	0.197	−0.559	−0.147	−18.22	−0.196	−4.67	0.706	−10.12

The coefficients of the equations (22) calculated from elements of Table III are $1.08\Delta\xi_1 - 2.36\Delta\xi_2 = 0.169$ d $-2.36\Delta\xi_1 + 1.08\Delta\xi_2 = -0.233$, giving as a final re- t $\xi_1 = 0.36$ and $\xi_2 = 0.21$ or $\alpha_c = 2/3$ and $L = 0.73\ R^2C$. r this optimum circuit, $I = 0.0956$. The circuit re- onse is given in Fig. 2 and compared with the conven- nal pi section derived from filter considerations.[14] The ler of magnitude of the error terms is

$$\epsilon_1 \simeq 0.001, \quad \epsilon_2 \simeq 0.0009, \quad \text{and} \quad \epsilon_3 \simeq 0.0008.$$

the event that the error terms should be too large, it uld be necessary to use new starting values of ξ_1 and and to repeat the procedure. The results of this par- ular calculation indicate a pronounced improvement m using this video coupling network when the driving e output capacitance exceeds the driven tube input pacitance by a substantial amount; the optimum ratio two to one might readily be realized with some tube es.

There are many cases in which the transient response of an amplifier is not a sole performance criterion. Other factors of interest, such as noise acceptance and adjacent channel rejection, can be taken into account by a modification of the basic method.

The "noise acceptance" of an amplifier may be defined by:

$$M = \int_0^\infty A^2(f)df,$$

in which $A(f)$ is the amplifier amplitude response at a frequency f. It can also be expressed as

$$M = \int_0^\infty [f_2'(t)]^2 dt,$$

where $f_2'(t)$ is the response to a δ function. M is proportional to the power of the noise present at the output of the amplifier, when its input is connected to a random (or "white") noise source.

[14] H. A. Wheeler, "Wide-band amplifiers for television," Proc. .E., vol. 27, pp. 429–437; July, 1939.

[15] V. D. Landon, "The band-pass low-pass analogy," Proc. I.R.E., vol. 24, pp. 1583–1584; December, 1936.

[16] P. R. Aigrain, B. R. Teare, Jr., and E. M. Williams, "Generalized theory of the band-pass low-pass analogy," to appear in Proc. I.R.E.

It would be almost hopelessly complex to try to choose the noise acceptance, and then minimize the transient distortion; indeed, the noise acceptance is given by a complex function of the circuit parameters. A preferable procedure is to minimize an expression of the form:

$$U = I + aM,$$

where a is some constant determined by the relative importance assigned to the transient response and to the noise acceptance. This procedure is particularly simple if one uses as test signal a delta-function. Corresponding to (7):

$$U = (1 + a)\int_0^\infty f_2'(t)dt - 2f_2'(T_0).$$

The calculations are similar to those involving transient response only, and the resulting equations are the same except that γ_n is replaced by $B_n/1+a$.

Adjacent channel rejection is of interest particularly in band-pass amplifiers in which it may be specified that some particular adjacent frequency must be strongly attenuated. As in the case of noise acceptance, it would be impractical to introduce this condition as a relation between the circuit parameters. It is, however, possible to handle this problem by minimizing the sum of the transient distortion and of the amplifier response to the undesired frequency. The amplifier input signal may be treated as the sum of a step function, for example, and a damped sine wave of frequency equal to the undesired frequency and with damping and amplitude dependent on the desired rejection. The frequency spectrum of such a signal appears in Fig. 3. Due to this additional

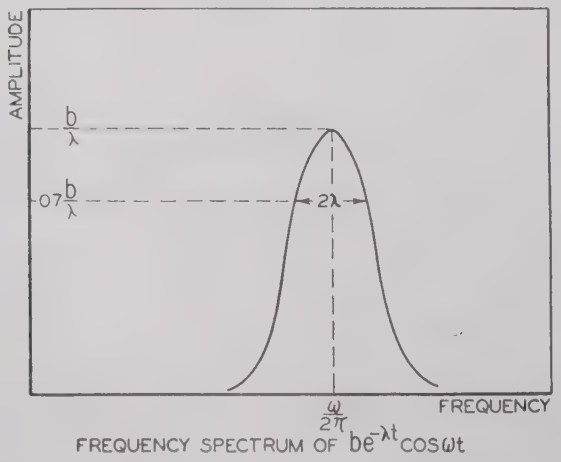

Fig. 3—Frequency spectrum of $be^{-\lambda t}\cos 2\pi ft$, the function used in minimizing adjacent channel rejection.

input signal, I as calculated from (1) will be increased by an amount which will be least if the desired frequency is properly rejected. If the damping and amplitude are small, the main part of the transient will be practically unaffected and the increase of I is given approximately by:

$$U_2 = I + \frac{A_f^2 b^2}{4\lambda}$$

A_f = response of amplifier at the rejected frequency

if the added function is $be^{-\lambda t}\cos 2\pi ft$. This procedure thus equivalent to minimizing the sum of the transien distortion and the amplifier response at the undesire frequency.

V. Cascade Amplifiers

The method of optimizing transient response is, course, applicable to cascade amplifiers. It is of consi erable interest to determine whether all stages should b identical for optimum transient response. If one cal $\xi_{11}, \xi_{12}, \xi_{21}, \cdots, \xi_{nm} \cdots$ the adjustable parameters r ferring to the 1st, 2nd, etc., circuit of the cascade, it evident that I can depend on these parameters onl through symmetrical combinations of $\xi_{11}, \xi_{12}, \cdots$, an $\xi_{21}, \xi_{22}, \cdots$. But any symmetrical function H of thes is such that

$$\frac{\partial H}{\partial \xi_{ik}}\left[\sum_R \xi_{1k} = \text{constant}\right] = 0 \quad \text{if} \quad \xi_{11} = \xi_{12} = \xi_{13},$$

i.e., any symmetrical function of a number of variable does not change in value when all the variables are ap proximately equal, if any two variables are changed b a small amount, the sum of all variables staying con stant.

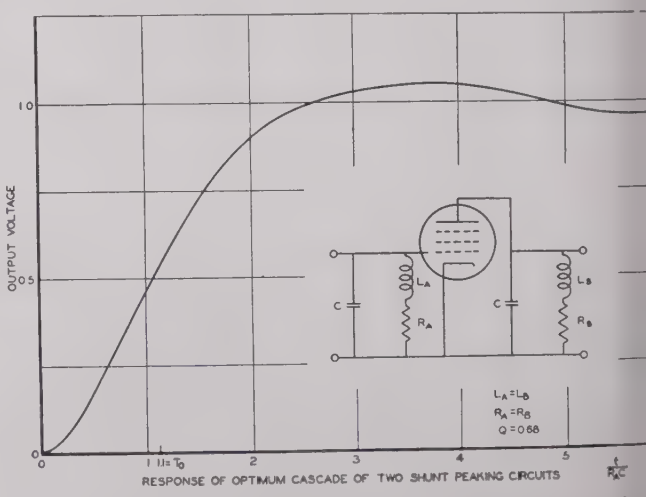

Fig. 4—Optimum cascade of two identical shunt-peaked circuits and its response.

Thus

$$\frac{\partial I}{\partial \xi_{1k}}\left[\sum_R \xi_{1k} = \text{constant}\right] = 0 \quad \text{for} \quad \xi_{11} = \xi_{12} = \xi_{13}$$

since I depends only on functions such as H.

The design for which $\xi_{12} = \xi_{12} = 0 \cdots$ is thus optimum provided that

$$\frac{\partial I}{\partial \xi_{11}} = 0, \qquad \frac{\partial I}{\partial \xi_{21}} = 0,$$

ıce the other optimizing relations will follow immediely from these. The consequence of this can be exessed in other words by the theorem "For optimum ınsient response in a video amplifier cascade,[17] all secons of the cascade should be made identical."

[17] These conclusions also apply to the video equivalent, when such sts, of a band-pass amplifier. This includes the case of staggered intermediate-frequency amplifiers, for instance, in which re are an even number of stages, each pair being symmetrically ed with respect to carrier frequency.

It is not possible to prove, but there are reasons to believe, that the optimum design for a section in a cascade is not very different from the optimum design of this section used alone. Indeed the amplitude as well as the phase distortion of each section will be cumulative, and thus it appears that the only way to improve the response of the amplifier as a whole is to make each section as good as possible. The results of a calculation of the optimum design for two shunt-peaking circuits are given in Fig. 4 and support this view.

Admittance of the 1B25 Microwave Switching Tube*

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Summary—**The operation of an RCA-1B25 neon-argon-filled w tube as a switching element in a transmission-line network at ra-high frequencies has been investigated. A test method has been veloped and theory formulated for evaluating the behavior of the e in terms of an equivalent susceptance and conductance. Experintal measurements have been made which show the variation he admittance characteristics as a function of gas pressure and of direct current.**

THE 1B25 IS a neon-argon-filled diode which fits into a coaxial transmission line and is used to switch a small ultra-high-frequency signal. Admittance characteristics of the tube were studied because of eir importance in tube replacement.

A block diagram of the test equipment used to mease the conductance and susceptance of the tubes under test is given in Fig. 1. The 1B25 tube terminates the 150-ohm tunable coaxial line, as indicated. The method of determining the admittance of the 1B25 in operation requires measurement of the voltages at junction k (Fig. 1) with the switching tube unfired and fired as a function of the electrical angular distance of the 1B25 from the junction. The ratio of these voltages is termed the switching ratio

$$SR = \left| V_{ku}/V_{kf} \right|, \tag{1}$$

wherein subscripts u and f refer to the unfired and fired states, respectively. This switching ratio was evaluated because of its applicability to operating conditions in the direction-finding equipment in which the tube finds its principal use.

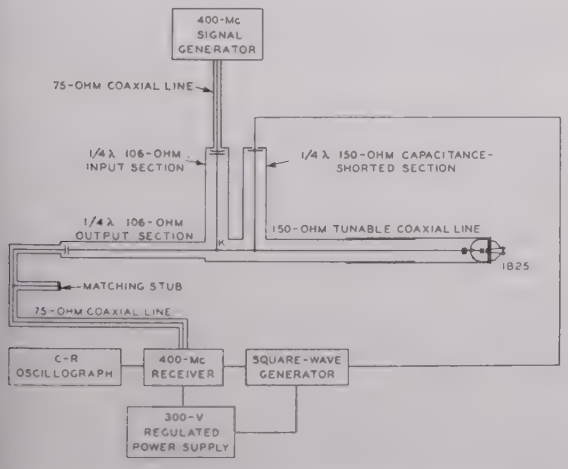

Fig. 1—Block diagram of experimental equipment.

The tube under test is fired at audio frequencies by a square wave of 250 peak volts. Included in the circuit is a frequency-doubler arrangement which delivers a square wave at twice the fundamental frequency. This double-frequency wave is applied to disable the receiver during part of each fired and unfired period. The switching ratio may then be measured as the relative height of two square-wave pips on the oscillograph.

The relative admittance of the 1B25 may be defined as

$$Y_u/Y_0 = jb_u + g_u, \text{ for the unfired state,} \tag{2}$$

and

$$Y_f/Y_0 = jb_f + g_f, \text{ for the fired state.} \tag{3}$$

Y_0 is the surge admittance of the line, Y_u and Y_f are the unfired and fired admittances of the 1B25, and the b's and g's represent corresponding relative susceptances and conductances.

The tunable coaxial line terminated by the 1B25 may be considered to have an electrical length of θ degrees

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from the junction k. When an analysis is made of the junction voltages for the fired and unfired condition, the following expression for switching ratio may be derived:

$$(SR)^2 = \frac{\left[3 + \dfrac{(g_f + 1)^2 + 2g_f + b_f^2}{(\cos\theta - b_f \sin\theta)^2 + (g_f \sin\theta)^2}\right]}{\left[3 + \dfrac{(g_u + 1)^2 + 2g_u + b_u^2}{(\cos\theta - b_u \sin\theta)^2 + (g_u \sin\theta)^2}\right]}. \quad (4)$$

This working expression gives the ratio squared of the magnitudes of the received signal for the two conditions of the tube as a function of the length of the tunable line and of a set of constants which represent the lumped relative admittance of the switching tube during nonconducting and conducting periods.

The results of various possible switching-tube admittances, plotted from (4), are illustrated in Fig. 2. The parameters chosen and the equivalent circuits representing the types of "short" introduced by the firing of the 1B25 are indicated in the accompanying legend. Very distinct characteristics classify the possible types of tube behavior.

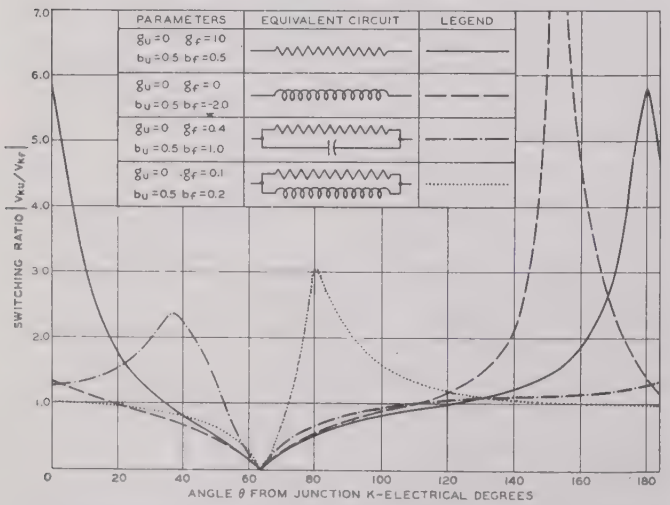


Fig. 2—Variation of switching ratio with length of tunable line, based on (4) for different possibilities of fired-tube admittance.

The inverse problem of determining the admittance characteristics from the experimentally measurable SR function required a graphical solution. Fig. 3 shows loci of $SR(\theta_{\text{min. fired}})$ and $\theta_{\text{min. fired}}$. The latter parameter was chosen as most convenient for the graphical analysis; it represents the electrical angular length of the tunable line for the condition that the junction voltage be a minimum during the fired condition. The switching ratio for $\theta_{\text{min. fired}}$ is in general nearly equal to the maximum switching ratio. In Fig. 3, g_u is assumed to be zero, as may be expected in a properly insulated tube. The value of b_u is 0.36, which corresponds to a $\theta_{\text{min. fired}}$ of 70.3°, or a capacitive reactance of 417 ohms at 400 Mc, and is the value to which all tubes are adjusted during manufacture. The intersection of loci $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$ defines b_f and g_f.

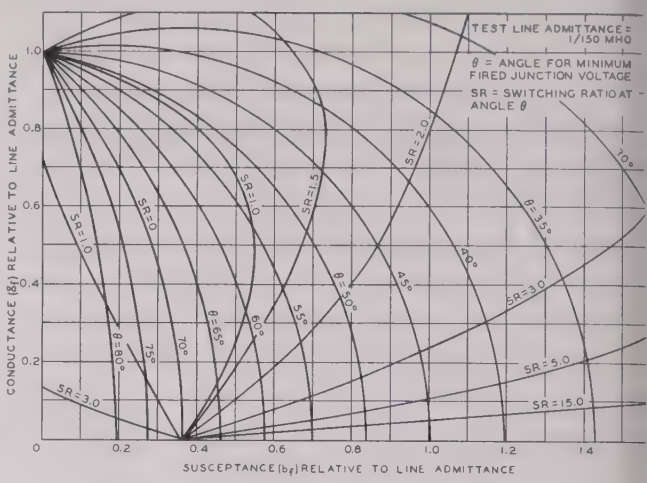

TEST LINE ADMITTANCE = 1/150 MHO
θ = ANGLE FOR MINIMUM FIRED JUNCTION VOLTAGE
SR = SWITCHING RATIO AT ANGLE θ
CONDUCTANCE (g_f) RELATIVE TO LINE ADMITTANCE
SUSCEPTANCE (b_f) RELATIVE TO LINE ADMITTANCE

Fig. 3—Chart for obtaining conductance and susceptance from $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$.

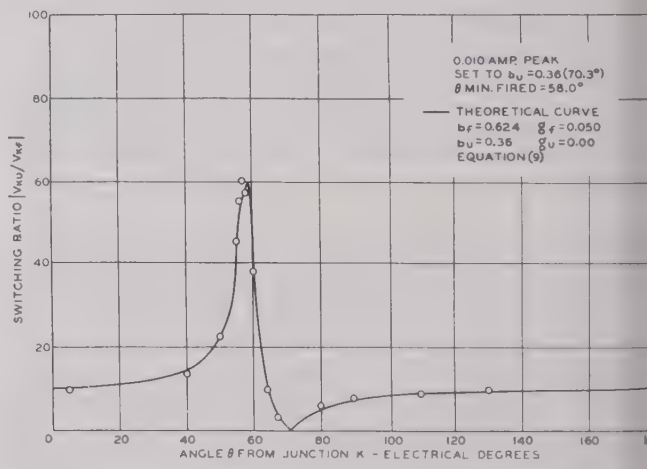

0.010 AMP. PEAK
SET TO b_u = 0.36 (70.3°)
θ MIN. FIRED = 58.0°
THEORETICAL CURVE
b_f = 0.624 g_f = 0.050
b_u = 0.36 g_u = 0.00
EQUATION (9)
SWITCHING RATIO $|V_{KU}/V_{KF}|$
ANGLE θ FROM JUNCTION K - ELECTRICAL DEGREES

Fig. 4—Variation of switching ratio with length of tunable line for typical 1B25, illustrating the accuracy with which the tube's behavior may be identified with admittance parameters.

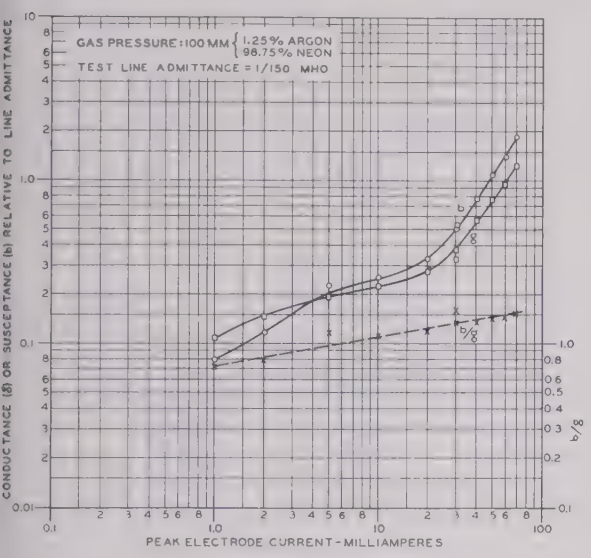
Experimental points for an $SR(\theta)$ curve are shown in Fig. 4. The solid line shows the theoretical SR curve according to (4) with the parameters b_f and g_f as obtained from $\theta_{\text{min. fired}}$ and $SR(\theta_{\text{min. fired}})$ and the graph of Fig. 3. The fired gap admittance is, thus, equivalent to a capacitive susceptance in parallel with a conductance. All of the 1B25 tubes measured exhibited a similar admittance at 400 Mc within the range of attainable currents.

A more fundamental method of showing the behavior of a 1B25 tube is to plot susceptance and conductance of the ionized gas as a function of the peak discharge current. It is convenient to define the relative susceptance and conductance of the ionized gas:

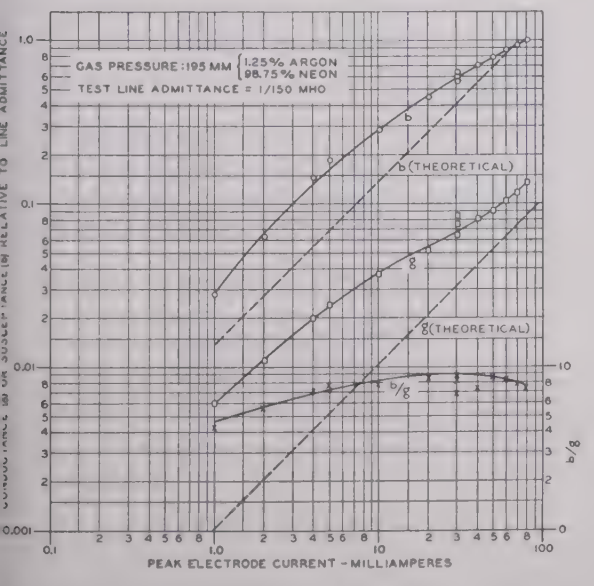
$$b = b_f - b_u \quad (5)$$

$$g = g_f - g_u.$$

gs. 5 and 6 show values of b, g, and the ratio of b/g as unction of peak current through the tube. Gas pres-e in these two cases was 100 and 195 mm of mercury a mixture of neon and $1\frac{1}{4}$ per cent argon. The standard 25 tube is filled to a pressure of 150 mm of mercury. rge variations of these characteristics were found, but


GAS PRESSURE: 100 MM { 1.25% ARGON 98.75% NEON
TEST LINE ADMITTANCE = 1/150 MHO
CONDUCTANCE (g) OR SUSCEPTANCE (b) RELATIVE TO LINE ADMITTANCE
b/g
PEAK ELECTRODE CURRENT - MILLIAMPERES

5—Variation of conductance and susceptance at 400 Mc as a unction of the direct current through a 1B25 tube having low gas pressure.


GAS PRESSURE: 195 MM { 1.25% ARGON 98.75% NEON
TEST LINE ADMITTANCE = 1/150 MHO
b (THEORETICAL)
g (THEORETICAL)
b/g
PEAK ELECTRODE CURRENT - MILLIAMPERES

6—Variation of conductance and susceptance at 400 Mc as a unction of the direct current through a 1B25 tube having high gas pressure.

ceptance and conductance always increased with the ct current. The admittance of a tube usually changes ing operation, particularly at the outset; the plotted ues were measured, therefore, after the tube had n aged at least 30 minutes. The differences between operation at 100 and 195 mm of mercury, as illustrated in Figs. 5 and 6, are quite typical of the general shift in characteristics with gas pressure. Conductance (particularly) and susceptance both increase with decrease in pressure.

In Fig. 6 theoretical curves also are shown for b and g variation with current. Since the theoretical computation is lengthy and necessarily approximate, it has been omitted. The calculated behavior agrees with experiment only to the extent of indicating trends.

A large switching ratio was desirable in the antenna-switching application. This ratio could be obtained at high gas-discharge currents, but practical limitations on tube life limited the current to about 40 ma. Operation at pressures lower than 150 mm was excluded because of excessive sputtering and eventual short-circuiting of the electrodes. An attempt was made to increase the switching efficiency by mechanically restricting the area of glow between the two electrodes. A ceramic collar was built to enclose the sides of the cathode as illustrated in the sketch of Fig. 7. This collar restricted the cathode-glow region to the flat surface facing the anode. A more uniform distribution of cathode glow resulted, which tended to prevent erratic shifts in the switching ratio.

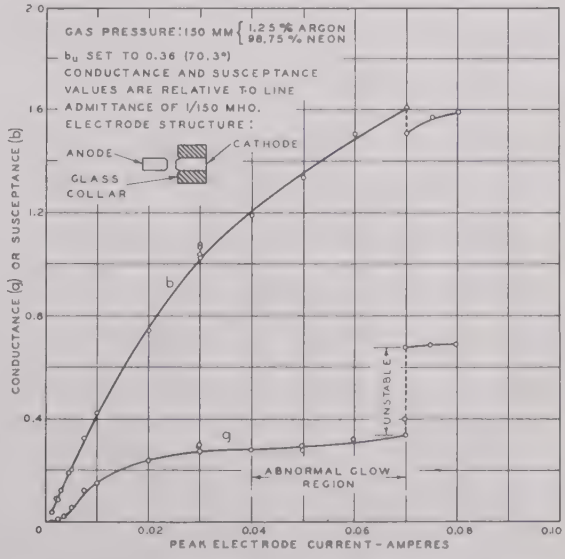

GAS PRESSURE: 150 MM { 1.25% ARGON 98.75% NEON
b_u SET TO 0.36 (70.3°)
CONDUCTANCE AND SUSCEPTANCE VALUES ARE RELATIVE TO LINE ADMITTANCE OF 1/150 MHO.
ELECTRODE STRUCTURE:
ANODE
CATHODE
GLASS COLLAR
CONDUCTANCE (g) OR SUSCEPTANCE (b)
UNSTABLE
ABNORMAL GLOW REGION
PEAK ELECTRODE CURRENT - AMPERES

Fig. 7—Variation of the 400-Mc conductance and susceptance with current for a restricted discharge.

The voltage drop was increased, however, due to the abnormal glow condition. Conductance and susceptance values for the restricted discharge are plotted in Fig. 7. At high current values beyond the indicated unstable point, the cathode glow jumped around the collar and returned to a condition of unrestricted discharge. It may be observed that restriction of the discharge increases the susceptance and decreases the conductance of the ionized region.

Bridged Reactance-Resistance Networks*

G. R. HARRIS†

Summary—**Consideration is given to the various possible six-arm, six-element *RC* bridged networks of which the parallel-tee is the previously known example. It is found that six symmetrical *RC* structures exist having the infinite attenuation property of the parallel-tee and other properties differing from those of this structure. The duality of certain pairs of such structures is demonstrated.**

Scope of This Paper

THIS PAPER is limited to the development of certain networks having six arms only, each arm comprising a single resistance or reactance of common sign. The chief property of interest of these networks is the infinite attenuation attained at some real frequency determined by the network constants.

The designation "bridged" for this class of networks is descriptive of less than half the number found, but is adopted by analogy with the familiar bridged-tee configuration, which is at the same time a pi in series with a single shunt arm.

The Problem

A familiar example of the above-defined class is the parallel-tee configuration of resistances and reactances, usually capacitive. This structure has been employed for a variety of purposes: as a power-supply filter,[1] a radio-frequency bridge,[2] a frequency-determining element of an oscillator,[3] a selective amplifier,[4] low- and high-pass filters,[5] a linear frequency discriminator,[6] a motor control circuit,[7] and others.

In view of this fairly extensive application, extending over a period of ten years or more it is of interest to inquire whether the parallel-tee configuration is the only three-terminal configuration of resistances and capacitances (or inductances) capable of affording infinite suppression at a selected frequency. It will be shown that the parallel-tee is not in fact unique, but is a member of a class of six six-arm structures having this property.

Development of Equivalent Networks

The parallel-tee configuration, including its two terminating impedances, can be drawn in generalized form as the eight-arm mesh of Fig. 1. This mesh can b considered either as a configuration of impedances c admittances; it happens that the equation stating th

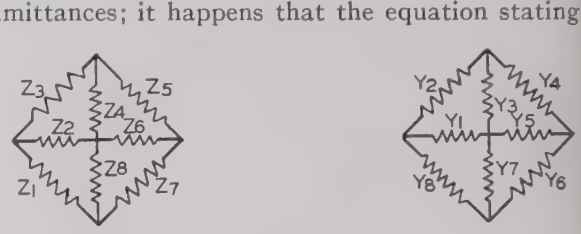


Fig. 1

conditions for infinite attenuation of the parallel-te falls in a more elegant form when expressed in admi tances. For half of the networks to be developed, how ever, the reverse is true, and thus it is convenient to b gin with both forms of the mesh in Fig. 1. It will b noted that the subscripts of the chord impedances Fig. 1(a) are those of the branch admittances of Fi 1(b), and vice versa. This inversion brings the equatio for dual networks into parallel form and halves th number of equations to be solved.

If the opposite branches Y_1 and Y_5 of Fig. 1(b) a made the terminating impedances, the parallel-tee co figuration of Fig. 2 appears. If, however, opposi chords are made terminating impedances, as Z_1 and of Fig. 1(a), the configuration of Fig. 3 is obtaine These two networks are dual, as the series and shu arms of either one appear as the correspondingly nu bered shunt and series arms, respectively, of the othe The configuration of Fig. 3 might well be called a serie pi network by analogy with the parallel-tee of Fig. 2.

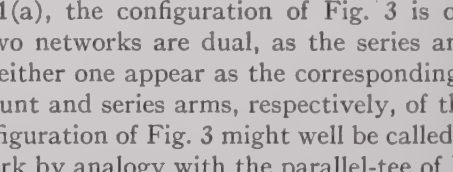


Fig. 2

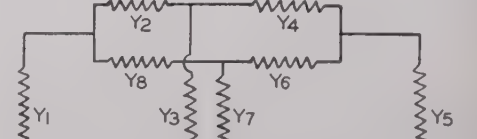


Fig. 3

Two other symmetrical configurations may be ob tained by making adjacent chords of Fig. 1(b) an adjacent branches of Fig. 1(a) the terminating im pedances. If Y_2 and Y_8 are so chosen in the first of thes cases, the configuration of Fig. 4 results. This may b called a shunt-ladder network, since it is in fact a ladd

* Decimal classification: R143. Original manuscript received by the Institute, October 29, 1948; revised manuscript received, February 23, 1949.

† Jones & Laughlin Steel Corporation, Pittsburgh, Pa.

[1] H. W. Augustadt, United States Patent No. 2,106,785; 1938.

[2] W. N. Tuttle, "Bridged-T and parallel-T null circuits for measurements at radio frequencies," Proc. I.R.E., vol. 28, pp. 23–30; January, 1940.

[3] W. G. Shepherd and R. O. Wise, "Variable-frequency bridge-type frequency-stabilized oscillators," Proc. I.R.E., vol. 31, pp. 256–269; June, 1943.

[4] H. H. Scott, "A new type of selective circuit and some applications," Proc. I.R.E., vol. 26, pp. 226–236; February, 1938.

[5] G. J. Thiessen, "R-C filter circuits," *Jour. Acous. Soc. Amer.*, vol. 16, pp. 275–279; April, 1945.

[6] J. R. Tillman, "Linear frequency discriminator," *Wireless Eng.*, vol. 23, pp. 281–286; October, 1946.

[7] D. S. Bond, United States Patent No. 2,429,257; 1947.

eries with a single shunt arm. If Z_2 and Z_8 are chosen erminating impedances, the configuration of Fig. 5 ears, which may be called a bridged-ladder. The net-ks of Figs. 4 and 5 are dual.

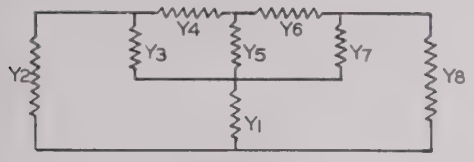


Fig. 4

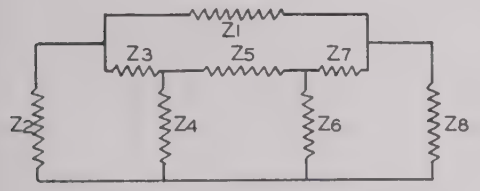


Fig. 5

Conditions for Balance

All four of the networks developed above, if considered as symmetrical about a vertical center line, have, of course, lattice equivalents, which may be easily derived through Bartlett's bisection theorem. It is, therefore, convenient and not inaccurate to speak of the conditions for infinite attenuation as the balance conditions.

Balance conditions may be determined by setting up the general circuit equations in terms of E, I, and either Y or Z, and equating either E or I to zero. As the network of Fig. 3 is drawn as a configuration of impedances, it will be convenient to determine its balance conditions from its mesh equations. The order of the meshes is obvious from Fig. 1(a) when E, the applied voltage, is considered to be in series with Z_1. The mesh currents are taken as flowing clockwise in each mesh.

$$
\begin{aligned}
E &= I_1(Z_1+Z_2+Z_3) - I_2Z_2 & 0 & & - I_4Z_8\\
0 &= - I_1Z_2 & + I_2(Z_2+Z_3+Z_4) & - I_3Z_4 & 0\\
0 &= 0 & - I_2Z_4 & + I_3(Z_4+Z_5+Z_6) & - I_4Z_6\\
0 &= - I_1Z_8 & 0 & - I_3Z_6 & + I_4(Z_6+Z_7+Z_8)
\end{aligned}
$$

$$
I_3 = E \qquad \begin{vmatrix} -Z_2 & Z_2+Z_3+Z_4 & 0 \\ 0 & -Z_4 & -Z_6 \\ -Z_8 & 0 & Z_6+Z_7+Z_8 \end{vmatrix} \Big/ \Delta
$$

he graphic method described by Tellegen[8] may be d to demonstrate the duality of these two pairs of figurations. If the configuration of Fig. 4, for exam-, is drawn in solid lines as in Fig. 6; a node placed in h of the four meshes of this configuration, as well as

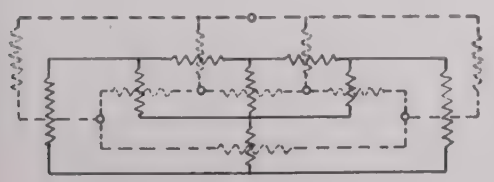


Fig. 6

outside corresponding to the mesh circumscribing configuration; and branches drawn from node to le so that each node is the junction of as many nches as its corresponding mesh has chords, then the ted-line configuration of Fig. 6 is obtained. The ted-line configuration is the dual of the solid-line figuration, and is seen to be the configuration of . 5 previously developed.

[8] B. D. H. Tellegen, "Geometrical configurations and duality of trical networks," *Phil. Tech. Rev.*, vol. 5, pp. 324–330; November,).

where Δ is the full determinant. The requirement for balance is that $I_3=0$. If this substitution is made in the above expression, the minor is easily expanded to:

$$0 = Z_2Z_4(Z_6+Z_7+Z_8) + Z_6Z_8(Z_2+Z_3+Z_4) \quad (1)$$

which expresses the conditions of balance for the network of Fig. 3. If the Z's in (1) are replaced by Y's of the same subscripts, the resulting equation in Y's expresses the conditions of balance for the parallel-tee network of Fig. 2.

The balance equation for the networks of Figs. 4 and 5 is derived in the same manner, and is found to be, for the shunt-ladder configuration of Fig. 4:

$$0 = Y_3Y_7(Y_4+Y_5+Y_6) + Y_5(Y_3Y_6+Y_4Y_7) + Y_4Y_6(Y_1+Y_3+Y_5+Y_7). \quad (2)$$

This equation in Z's applies to the bridged-ladder configuration of Fig. 5.

Physically Realizable Structures: Parallel-tee and Series-Pi

As has been mentioned, this paper is confined to six-arm structures of six elements only. Since RC networks are of more general application than RL networks, the latter will not be discussed, although it is obvious that

for every *RC* structure considered, a corresponding *RL* structure exists.

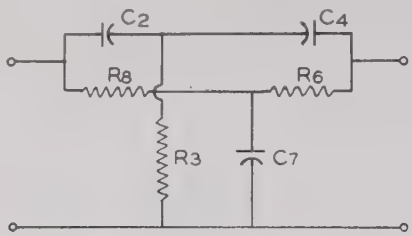


Fig. 7

The *RC* parallel-tee structure corresponding to Fig. 2 is that of Fig. 7. Its balance conditions are expressed in general terms by the *Y* form of (1) above

$$0 = Y_2Y_4(Y_6 + Y_7 + Y_8) + Y_6Y_8(Y_2 + Y_3 + Y_4).$$

Substituting g_3, g_6, g_8, jb_2, jb_4, and jb_7 for the corresponding *Y*'s gives:

$$0 = -b_2b_4(g_6 + g_8 + jb_7) + g_6g_8(g_3 + j[b_2 + b_4].)$$

The real and imaginary terms may be separated and equated to zero in the usual manner:

Real: $$g_3g_6g_8 = b_2b_4(g_6 + g_8) \tag{3}$$

Imaginary: $$b_2b_4b_7 = (b_2 + b_4)g_6g_8. \tag{4}$$

At this point it is convenient to replace the *g*'s and *b*'s by the *R*'s and *C*'s of Fig. 7:

$$w^2C_2C_4\left(\frac{1}{R_6} + \frac{1}{R_8}\right) = \frac{1}{R_3R_6R_8}$$

$$w^3C_2C_4C_7 = \frac{w(C_2 + C_4)}{R_6R_8}$$

$$w^2 = \frac{1}{C_2C_4R_3(R_6 + R_8)} \tag{5}$$

$$w^2 = \frac{C_2 + C_4}{C_2C_4C_7R_6R_8} \tag{6}$$

$$f = \frac{1}{2\pi\sqrt{C_2C_4R_3(R_6 + R_8)}} \tag{7}$$

$$f = \frac{1}{2\pi}\sqrt{\frac{C_2 + C_4}{C_2C_4C_7R_6R_8}}. \tag{8}$$

If (5) is equated to (6),

$$(C_2 + C_4)C_2C_4R_3(R_6 + R_8) = C_2C_4C_7R_6R_8$$

$$\frac{C_2 + C_4}{C_7} = \frac{R_6R_8}{R_3(R_6 + R_8)}. \tag{9}$$

The balance conditions for the structure of Fig. 7 may thus be expressed by two equations, either (7) or (8), and (9), the latter being independent of frequency.

The *RC* structure corresponding to the configuration of Fig. 3 is that of Fig. 8, the dual of Fig. 7. It is, of course, unnecessary to derive its balance equations, as they may be obtained from (5) through (9) previously derived merely by replacing all *C*'s with *R*'s of corresponding subscripts, and vice versa. These corresponding equations are:

Imaginary: $$w^2 = R_2R_4C_3(C_6 + C_8) \tag{10}$$

Real: $$w^2 = \frac{R_2 + R_4}{R_2R_4R_7C_6C_8} \tag{11}$$

$$= \frac{1}{2\pi\sqrt{R_2R_4C_3(C_6 + C_8)}} \tag{12}$$

$$f = \frac{1}{2\pi}\sqrt{\frac{R_2 + R_4}{R_2R_4R_7C_6C_8}} \tag{13}$$

$$\frac{R_2 + R_4}{R_7} = \frac{C_6C_8}{C_3(C_6 + C_8)}. \tag{14}$$

Either of (12) or (13) together with (14) express the balance conditions for the structure of Fig. 8.

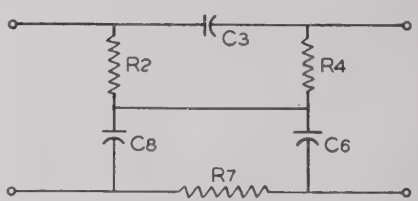


Fig. 8

Physically Realizable Structures: Shunt-Ladder and Bridged-Ladder

The parallel-tee *RC* structure discussed above wa previously known, and the series-pi is its dual. Th problem of deriving real *RC* structures for the ne works of Figs. 4 and 5 however, is less easy of approac Analogy with the previously developed structures an regard for symmetry suggest that if the three shun arms, Y_3, Y_5, and Y_7 of Fig. 4, are made capacitance the remaining arms must be resistances and the stru ture of Fig. 9 is thus obtained. Its balance equatio

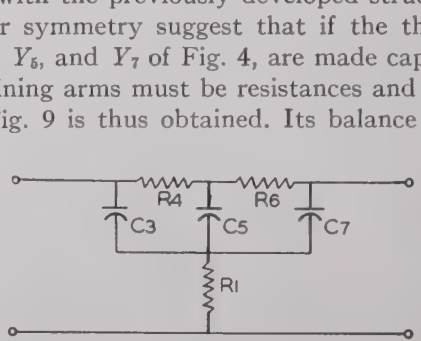


Fig. 9

may be derived from (2) by substituting the conduc ances and susceptances of Fig. 9 for the generalized a mittances:

$$0 = -b_3b_7(g_4 + g_6 + jb_5) + jb_5(g_6jb_3 + g_4jb_7) + g_4g_6(g_1 + j[b_3 + b_5 + b_7]).$$

The real and imaginary terms of the above may separated and equated to zero:

Real: $$g_1g_4g_6 - b_3b_7(g_4 + g_6) - b_5(g_6b_3 + g_4b_7) =$$

Imaginary: $$b_3b_5b_7 - g_4g_6(b_3 + b_5 + b_7) =$$

terms of the element designations of Fig. 9 these uations become:

al: $1 = R_1w^2(R_4C_3[C_7 + C_5] + R_6C_7[C_3 + C_5])$ (15)

aginary: $w^2 = \dfrac{C_3 + C_5 + C_7}{R_4R_6C_3C_5C_7}$ (16)

$$f = \frac{1}{2\pi}\sqrt{\frac{C_3 + C_5 + C_7}{R_4R_6C_3C_5C_7}}. \quad (17)$$

the value for w^2 of (16) is substituted in (15) the re-ting equation is:

$$\frac{R_4R_6C_3C_5C_7}{R_1(C_3+C_5+C_7)} = R_4C_3(C_7+C_5)+R_6C_7(C_3+C_5). \quad (18)$$

The balance conditions for the structure of Fig. 9 may is be expressed by (17) and (18), the first giving the quency at which balance occurs, and the second a ation which the network elements must satisfy.

Although the networks of Figs. 7 and 8 are the only ssible symmetrical *RC* structures corresponding to the nfigurations of Figs. 2 and 3, the structure of Fig. 9 is a unique embodiment of the configuration of Fig. 4. arms Y_3, Y_5, and Y_7 are made resistances and the re-ining arms capacitances, the structure of Fig. 10 is

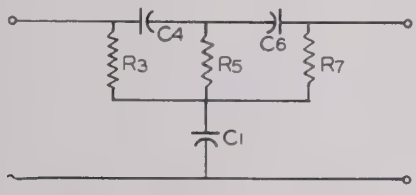


Fig. 10

ained. Its balance equations, obtained as before, are:

l: $1 = w^2C_4C_6(R_5R_7 + R_3R_7 + R_3R_5)$ (19)

ginary: $w^2C_1C_4C_6 = \dfrac{C_4(R_3 + R_5) + C_6(R_5 + R_7)}{R_3R_5R_7}$. (20)

m these are derived:

$$f = \frac{1}{2\pi\sqrt{C_4C_6(R_5R_7 + R_3R_7 + R_3R_5)}} \quad (21)$$

$$\frac{C_4(R_3+R_5)+C_6(R_5+R_7)}{C_1} = \frac{R_3R_5R_7}{R_5R_7+R_3R_7+R_3R_5}. \quad (22)$$

The bridged-ladder structures dual to the shunt-lad-structures of Figs. 9 and 10 may now be drawn im-

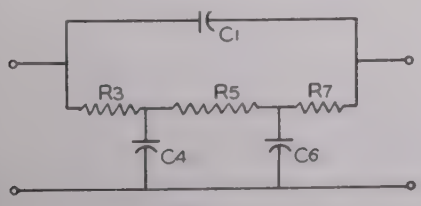


Fig. 11

mediately as Figs. 11 and 12, respectively, and their balance equations written down by inspection of those previously obtained. These equations for the structure of Fig. 11 are:

$$f = \frac{1}{2\pi}\sqrt{\frac{R_3 + R_5 + R_7}{C_4C_6R_3R\ R_7}} \quad (23)$$

$$\frac{C_4C_6R_3R_5R_7}{C_1(R_3+R_5+R_7)} = C_4R_3(R_7+R_5)+C_6R_7(R_3+R_5). \quad (24)$$

The equations of balance for the structure of Fig. 12 are:

$$f = \frac{1}{2\pi\sqrt{R_4R_6(C_5C_7 + C_3C_7 + C_3C_5)}} \quad (25)$$

$$\frac{R_4(C_3 + C_5) + R_6(C_5 + C_7)}{R_1} = \frac{C_3C_5C_7}{C_5C_7 + C_3C_7 + C_3C_5}. \quad (26)$$

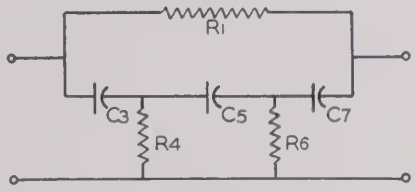


Fig. 12

Properties of Simplified Structures

When elements of the structures of Figs. 7–12, inclusive, are duplicated or triplicated, the balance equations reduce to more manageable forms, and, incidentally, reveal network properties which are masked by the general equations (7) through (26). If in the structures of Figs. 7 and 8 the elements having subscripts 2 and 6 are set equal to those of subscripts 4 and 8, respectively, the equations of balance become:

For Fig. 7:

$$f = \frac{1}{2\pi\sqrt{2}C_2\sqrt{R_3R_6}} \quad (27)$$

or

$$f = \frac{\sqrt{2}}{2\pi R_6\sqrt{C_2C_7}} \quad (28)$$

and

$$\frac{4C_2}{C_7} = \frac{R_6}{R_3}. \quad (29)$$

For Figure 8:

$$f = \frac{1}{2\pi\sqrt{2}R_2\sqrt{C_3C_6}} \quad (30)$$

or

$$f = \frac{\sqrt{2}}{2\pi C_6\sqrt{R_2R_7}} \quad (31)$$

and

$$\frac{4R_2}{R_7} = \frac{C_6}{C_3}. \tag{32}$$

As the properties of the parallel-tee have been discussed rather extensively and those of the series-pi are identical, no further consideration will be given them here.

If in the structures of Figs. 9–12, inclusive, elements of subscripts 3 and 4 are set equal to those of subscripts 7 and 6, respectively, the equations of balance become: For Fig. 9:

$$f = \frac{\sqrt{1 + 2\frac{C_3}{C_5}}}{2\pi C_3 R_4} \tag{33}$$

$$\frac{R_4}{R_1} = \frac{4C_3}{C_5} + \frac{2C_5}{C_3} + 6 \tag{34}$$

For Fig. 10:

$$f = \frac{1}{2\pi R_3 C_4 \sqrt{1 + 2\frac{R_5}{R_3}}} \tag{35}$$

$$\frac{C_1}{C_4} = \frac{4R_5}{R_3} + \frac{2R_3}{R_5} + 6. \tag{36}$$

For Fig. 11:

$$f = \frac{\sqrt{1 + \frac{2R_3}{R_5}}}{2\pi R_3 C_4} \tag{37}$$

$$\frac{C_4}{C_1} = \frac{4R_3}{R_5} + \frac{2R_5}{R_3} + 6. \tag{38}$$

For Fig. 12:

$$f = \frac{1}{2\pi C_3 R_4 \sqrt{1 + 2\frac{C_5}{C_3}}} \tag{39}$$

$$\frac{R_1}{R_4} = \frac{4C_5}{C_3} + \frac{2C_3}{C_5} + 6. \tag{40}$$

These equations are significantly different from the corresponding equations (27) through (32). The frequency-determining equations here contain a term which is a ratio of two like elements; this makes it possible to multiply or divide within certain limits the frequency determined by the duplicated elements. Consider, for example, (33) and (34), which apply to Fig. 9. If $C_3/C_5 = 1$, these become:

$$f = \frac{\sqrt{3}}{2\pi C_3 R_4} \tag{41}$$

$$\frac{R_4}{R_1} = 12. \tag{42}$$

Suppose, however, that $C_3/C_5 = 2$. Then

$$f = \frac{\sqrt{5}}{2\pi C_3 R_4} \tag{43}$$

$$\frac{R_4}{R_1} = 15. \tag{44}$$

A more interesting property appears from further consideration of these equations. Suppose $C_3/C_5 = 1/2$. Then

$$f = \frac{\sqrt{2}}{2\pi C_3 R_4} \tag{45}$$

$$\frac{R_4}{R_1} = 12. \tag{46}$$

For given values of C_3, R_4, and R_1 there are two independent frequencies of balance, (41) and (45), corresponding to two particular values of C_5. This double frequency of balance results from the form of (34) and corresponding equations for the other structures of this group.

In (34) for example, let $R_4/R_3 = P$, and $C_3/C_5 = Q$. The equation then becomes:

$$P = 4Q + \frac{2}{Q} + 6. \tag{47}$$

It has been shown just previously that $P = 12$ for two different values of Q. The minimum values of P and Q are obtained in the usual manner by differentiating P with respect to Q and setting the derivative equal to zero.

$$\frac{dP}{dQ} = 4 - \frac{2}{Q^2} = 0$$

$$Q^2 = \frac{1}{2}$$

$$Q_{\min} = \frac{1}{\sqrt{2}} \quad \text{or} \quad 0.707 \tag{48}$$

$$P_{\min} = \frac{4}{\sqrt{2}} + 2\sqrt{2} + 6 = 11.656. \tag{49}$$

Values of P less than this cannot be chosen if real results are to be obtained from (33) through (40). A graph of (47) is shown in Fig. 13, the Q scale being logarithmic. Drawn in this manner, the curve is a parabola.

The shunt-ladder and bridged-ladder structures differ from the parallel-tee and series-pi in still another way. It is possible to make either all R's or all C's the same value in both the latter two structures, but not in former structures. In the shunt-ladder and bridged-ladder configurations, the three resistances or capacitances occupying similar arms may be made equal, but not the elements of the other three arms. In Fig. 9, for example, all C's can be made equal, but not the R's; in

10 the R's can be made equal, but not the C's. This ows from (49) which shows that the minimum value ' is a little less than 12.

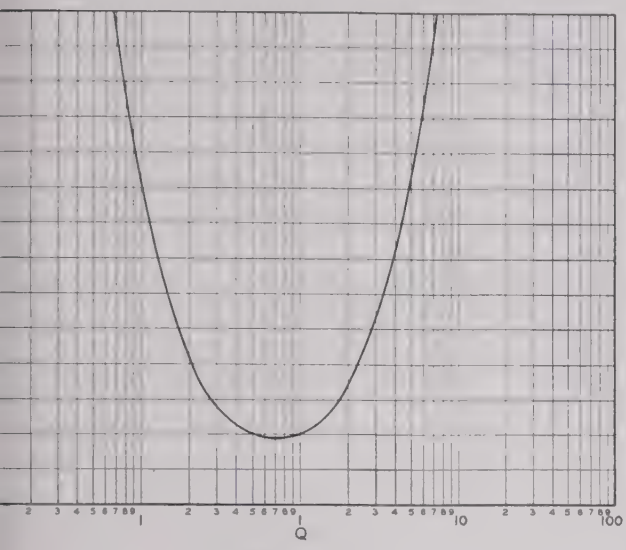


Fig. 13

lthough this paper will not discuss particular ap- ations of the structures considered, some practical antages of the shunt-ladder and bridged-ladder ctures will be briefly mentioned. Where networks balancing at low frequencies are desired, the structure of Fig. 12 offers economies in the size of the capacitances required. Where a high frequency balance is wanted, the structure of Fig. 9 is more suitable, as its capacitance values may be made appreciably larger for a given balance frequency than those of the others. The structure of Fig. 9 is useful if the three capacitances are to be varied, as they have a common terminal. This terminal need not be too far above ground potential, since the ratio of the series to the shunt resistances may be made fairly large.

Additional Networks

The configurations of Figs. 2 through 5 are not all those which may be developed from the mesh of Fig. 1. Two other such exist, but as they are not symmetrical and do not appear to possess any properties not previously known, they will be mentioned only briefly. One is obtained from making an adjoining branch and chord the terminating impedances; the resulting three-terminal configuration cannot be embodied in an RC structure which balances at any real frequency. The other is obtained from making an opposite branch and chord the terminating impedances; the resulting four-terminal configuration is that of the well-known Anderson Bridge. It is possible to balance this structure when only two of its six arms are capacitances, but in this respect it merely equals the Wien RC Bridge. Each of these additional configurations is its own dual.

Ionospheric Virtual Height Measurements at 100 Kilocycles*

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ummary—A simple high-power ionosphere sounding equipment se at low frequencies is described. Its application to the meas- ent of virtual height at vertical incidence at a carrier frequency 0 kc is reported. Results of nearly a year of intermittent night- measurements showed a remarkably large variation in virtual t which ranged from about 84 km to as high as 106 km. Some nce was found to indicate that frequently at night the reflecting n consists of clouds or patches of ionization, rather than the nearly uniform ionization characteristic of the regular layers gh frequencies. A rotation of the polarization of the reflected l with respect to that of the transmitted signal was observed.

I. Introduction

RECENTLY, considerable interest has arisen in the behavior of the ionosphere at low and very low frequencies. The development of low-fre- ncy precision aids to navigation, such as lf Loran, has necessitated a detailed knowledge of the variations in the electrical properties of the ionospheric reflecting region at these frequencies. Of particular interest is the virtual height[1] of reflection of low frequency sky-wave signals, for upon the virtual height depends the time delay and arrival angle of a sky wave. It is the purpose of this paper to report the initial results of virtual height measurements made at a carrier frequency of 100 kc, and to describe the special equipment developed for this purpose.

Much of the knowledge of the electrical and physical properties of the upper atmosphere has been obtained directly or indirectly by radio methods applied mainly at frequencies above 500 kc. The foremost of these is the group retardation or "pulse" method of Breit and Tuve which gives the virtual height of reflection of a short pulse at a given carrier frequency. Because the heights

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Stanford University, Stanford, Calif.

[1] "Virtual height" is defined as the height at which a signal would be reflected, if the ionosphere were replaced by a perfectly reflecting surface.

at which propagation is influenced by the presence of ionization tend to increase with the frequency, the characteristics of the lowest regions of the ionosphere can best be investigated by radio methods only at low frequencies. Thus it would seem that the study of the ionospheric reflections of low-frequency waves offers an excellent means of increasing our knowledge of the lower ionosphere.

A limited amount of data on the propagation of sky waves at very low frequencies have been obtained by the continuous-wave interference method[2] and by the observation of signals from lightning discharges.[3] In the continuous-wave technique, variations in effective height are determined from changes in the relative phase of ground and reflected waves arriving at a distant receiver. Polarization and field intensity can also be determined by this method. One of its limitations is the difficulty in interpreting the results when more than one reflected wave is present, which is the usual situation during nighttime. Another difficulty arises in attempting to apply the method at vertical incidence, where one must somehow attenuate the ground wave by several orders of magnitude.

The other method makes use of the fact that a lightning discharge radiates a pulse-type signal which is short enough to permit the resolution of multiple reflections between earth and ionosphere. The virtual height of the reflecting layer and the distance of the source are found from a measurement of the relative times of arrival of three or more pulses which have been reflected from the ionosphere a known number of times. This method is fundamentally restricted to very low frequencies, since the principal frequency components in a typical lightning discharge lie below about twenty kc. Various practical limitations prevent the fullest utilization of these natural radio signals. These include t difficulty of determining the location and exact cha acter of the discharge. Furthermore, there is eviden to indicate that abnormal ionization may be produc above a thunderstorm. If this be the case, then vertic incidence measurements of the normal ionosphere wou be impossible when using this method.

[2] J. E. Best, J. A. Ratcliffe, and M. V. Wilkes, "Experimental investigations of very long waves reflected from the ionosphere," *Proc. Roy. Soc.*, vol. 156, pp. 614–633; September, 1936.

[3] B. F. J. Schonland, et al., "The wave form of atmospherics at night," *Proc. Roy. Soc.*, vol. 176, pp. 180–202; October, 1940.

II. Apparatus

General Description

The transmitted pulse consists of an exponential damped oscillation which is produced about once eve second by directly exciting a large "inverted-L" type antenna. This pulse travels by way of the ground a the ionosphere to the receiver, located 2.7 km away. block diagram of the system is shown in Fig. 1. T ground and principal sky waves are radiated by t 30-meter high vertical part and the 440-meter lo

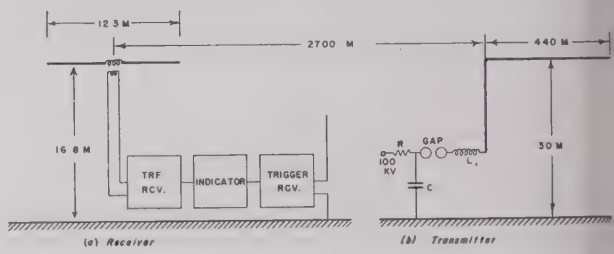


Fig. 1—Block diagram.

horizontal part of the antenna, respectively. Since t antenna is less than a quarter-wave long at 100 kc, loading inductance (L_1) is required to tune the anten to resonance. Energy for the pulse is obtained by slow charging a relatively large capacitor C, through a hig resistance R, from a dc source of high voltage rangi from 100 to 200 kv. The charged capacitor is connecte in series with the antenna by means of a variable-leng

Fig. 2—Transmitter site.

ere gap. One of the spheres is mounted on an arm ch rotates at a rate of about 1 rps. An instant before moving sphere reaches the position nearest the fixed ere, the gap breaks down and the energy stored in capacitor flows into the inductance and the antenna ch may be regarded simply as a transmission line. repeated reflections of diminishing amplitude pro- an exponentially damped oscillation whose period ends primarily on the line length and the size of L_1, whose time constant depends primarily on the re- ance of the ground connection. To reduce the pulse th, additional resistance may be placed in series the antenna. With typical circuit constants, no ex- al resistance, and an applied voltage of 100 kv, the k current in the pulse is 200 amperes, and the time stant is 150 μs. A photograph of the transmitting enna and the Ryan High Voltage Laboratory is wn in Fig. 2.

he principal receiving problem is obtaining an ade- te signal-to-noise ratio. In spite of the large current ing in the transmitting antenna, the power radiated the horizontal part is only of the order of 1 kw use of the extremely low radiation resistance. At frequencies and at low and medium latitudes, ospheric static is usually the limiting factor in re- er sensitivity. A substantial reduction in the pickup ich static is effected by use of a balanced horizontal dipole which is relatively insensitive to signals arriving at low vertical angles. To permit the "nulling out" of ground waves and to allow the polarization of the down-coming waves to be investigated, the dipole is universally mounted. Azimuth and tilt are controlled at the receiver by two sets of selsyns. By proper orientation of the dipole in the horizontal plane, its response to vertically polarized signals can be reduced by 60 db below that with the dipole vertical. The dipole is shown in the photograph of the receiving site in Fig. 3.

The pulses picked up by the dipole are amplified by a simple broadband TRF receiver of straightforward design. The output of the receiver is fed to a five-inch oscilloscope whose time base is triggered slightly in advance of the arrival of the ground pulse at the indicator by the strong ground pulse which is received on a broadly tuned vertical antenna. The time delay of the echoes with respect to the ground pulse is measured by means of timing markers which are derived from a 30-kc ringing oscillator. The ringing oscillator is triggered by the oscilloscope gate voltage. A photographic record of each trace is made with a motor-driven single-shot 16-mm movie camera, which is controlled by the gate voltage from the oscilloscope in such a way that the film moves ahead one frame each time the sweep is triggered. The time is indicated on each frame by a seconds-counter situated in the upper right-hand corner directly in front of the cathode-ray tube.

Power Supply

With the relatively low repetition rate of one pulse every one and one-half seconds, the average demand from the power supply is small, being about twenty watts under typical operating conditions. A maximum voltage of about 200 kv is obtained from nine cascaded half-wave rectifiers which are supplied from a 23-kv, 60-cycle transformer. A simplified schematic of the supply is shown in Fig. 4. Each stage consists of two 27,000-volt, 0.125-μf capacitors in series with a half-wave high-vacuum rectifier, type 250R. To avoid the problem of filament transformer insulation, the filament power is obtained from a 1-kw 450-kc oscillator which supplies

Fig. 3—Receiver site.

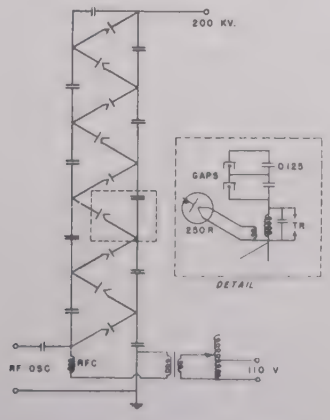


Fig. 4—Power supply schematic.

a current of one ampere to the autotransformer primaries which are connected in series through the capacitors. The secondaries supply the required filament current of ten amperes to each of the nine rectifiers.

Each autotransformer is tuned approximately to resonance by means of a shunt capacitance. The effect of variations in primary current from stage to stage is compensated by individually adjusting the tuning of each stage. The measurement of filament power is readily made with the aid of an optical pyrometer which is first calibrated, using a filament carrying a known effective current.

Over-voltage protection is provided by spark gaps connected across each capacitor, and "transmit-receive" gas tubes connected across each auto transformer.

Antenna Current

The current which flows in the antenna can be determined by considering the circuit to be made up of an inductance in series with a transmission line and a source of voltage. It is assumed that the capacitance C, which appears in the block diagram of Fig. 1, is so large compared with the antenna capacitance that the variation in voltage across C during the charging of the line is small and can be neglected to a good approximation.

The method of solution most adaptable to the present problem is to consider the total current to be the superposition of a series of sinusoidal components. Their amplitudes and frequencies may be determined by solving the transmission line differential equations in the presence of the given initial and boundary conditions by the method of the Laplace Transformation.[4] From such an analysis one finds the current at a distance x from the sending end of the line to be

$$I = \frac{4E}{\pi z_0} \sum_{n=1}^{\infty} \sin 2\pi f_n t \, \frac{f_0 \left[\cos \frac{\pi f_n}{2lf_0} x - \frac{2\pi f_n L_1}{z_0} \sin \frac{\pi f_n}{2lf_0} x \right]}{f_n \left[1 + 4f_0 \frac{L_1}{z_0} + 4\pi^2 f_n^2 \left(\frac{L_1}{z_0} \right)^2 \right]},$$

and the frequency of the nth mode to be given by

$$\tan \frac{\pi}{2} \frac{f_n}{f_0} = \frac{z_0}{2\pi f_n L_1} \qquad (2)$$

where

E = voltage of source capacitor
z_0 = characteristic impedance of line
f_n = frequency of the nth mode of oscillation
$f_0 = c/4l$ = natural resonant frequency of line
c = velocity of light
l = length of line
L_1 = loading inductance.

[4] H. S. Carslaw and J. C. Jaeger, "Operational Methods in Applied Mathematics," Oxford University Press, 1941.

Equation (1) is useful in obtaining the amplitud the fundamental component of current from which field intensity of the radiated wave may be calcula Equation (2), which is best solved graphically, gives loading inductance required to tune the line to the sired frequency.

Field Intensity

The complete expression for the radiation compone of the field produced by an "Γ" antenna is rather volved;[5] however, for vertical incidence work, the f intensity of interest is that directly above the anten This may be computed readily by assuming a sinusoi current distribution on the antenna, and adding direct and ground-reflected components produced the horizontal part. This leads to the following exp sion for the magnitude of the field intensity in volts meter at the receiving antenna assuming unity ref tion coefficient at the ionosphere:

$$E_r = \frac{30 I_0}{d} \frac{1 - \cos \frac{2\pi b}{\lambda}}{\sin 2\pi \frac{a+b}{\lambda}} \left| 1 - r_T \underline{/\alpha_T} \right| \left| 1 - r_R \underline{/\alpha_R} \right|,$$

where

I_0 = base current in amperes
d = total path distance in meters
b = length of transmitting antenna flat top meters
a = height of transmitting antenna above grou
λ = operating wavelength in meters
r_T, r_R = magnitudes of ground reflection coefficients transmitter and receiver, respectively (nor incidence, horizontal polarization)

$$\alpha_T = \rho_T + \frac{4\pi a}{\lambda}$$

$$\alpha_R = \rho_R + \frac{4\pi h}{\lambda}$$

ρ_T, ρ_R = phase angles of ground reflection coefficients transmitter and receiver, respectively (nor incidence, horizontal polar)
h = height of receiving antenna above ground.

As an illustration of the magnitude of the field int sity to be expected, it is found that in a typical case

[5] G. W. Pierce, "Electric Oscillations and Electric Waves," Graw-Hill Book Co., Inc., New York, N. Y., First Edition, 1 Chap. IX.

ich $I_0 = 302$ amperes and $d = 180$ km (round trip dis-ice to ionosphere), the peak field intensity at the re-ving dipole calculated from (3) is 340 μv per meter. is of interest to note that, for an antenna which is vated only a small fraction of a wavelength above ound, as is the case here, the upward field is increased preciably as the soil conductivity is reduced.

The practical measurement of the peak field intensity the reflected pulses may be carried out by calibrating e receiver in terms of the total field present at the tenna. This calibration is effected by using the ground lse as a standard field[6] in which the dipole is oriented a known angle, with respect to the electric vector of e ground wave.

Under favorable conditions, a loop antenna may be ed directly to measure the strength of the downcoming noes. Usually, however, the response of a loop to noise d interfering signals is too great to permit such easurement.

ceiver

The receiver used in these experiments is a simple RF type having three tuned circuits (not counting the ned antenna). The dipole transmission line is coupled the receiver through an electrostatically shielded nsformer, the over-all bandwidth of which is about kc.

The pulse response of the receiver is best described the oscillogram of Fig. 5, which shows the form of the -μs ground pulse as it appeared on the indicator. The

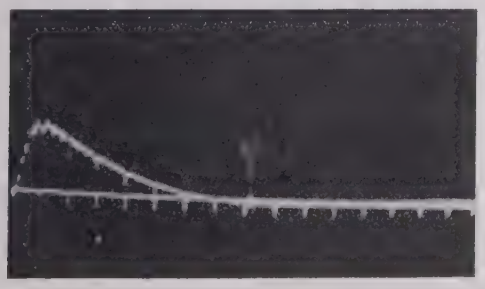


Fig. 5—Ground pulse, $T = 75$ μs, 5-km markers.

gative pips are timing markers spaced $33\frac{1}{3}$ μs apart. e pulse rise time is seen to be about 30 μs and the ration measured to one-third peak amplitude is about 0 μs, or twice the input time constant. The wideband the detector and following circuits permits some rip-e of twice the carrier frequency to appear on the out-t pulse. This ripple is actually quite useful, since it ves as a time scale for measuring pulse rise time more curately than is possible with the 5-km markers.

[6] The field intensity of the ground wave may be measured approximately with a loop-type intensity meter in which the induced se is matched by a similar inserted pulse of known peak amplitude. The approximation is that the effective height of the loop is umed the same for the sideband components of the pulse as for carrier, although it actually varies directly with frequency for a all loop.

III. Results

Virtual Heights

Apparatus of the type described in previous sections has been used to measure the virtual heights of the nighttime E layer of the ionosphere at a carrier frequency of 100 kc. These measurements, made on several occasions from the fall of 1947 to the summer of 1948, show certain interesting effects which will be described.

A rather unusual oscillogram was selected to illustrate the measurement technique, and is shown in Fig. 6. It is unusual in that it is easy to explain. The effective beginning of the ground pulse coincides with the start of the trace and the timing markers. The limiter in the receiver flattened the top of the ground pulse and thus prevented the otherwise excessively strong ground pulse from carrying the timing markers out of the picture. The virtual height or time delay of the first reflection is measured between corresponding parts of the ground

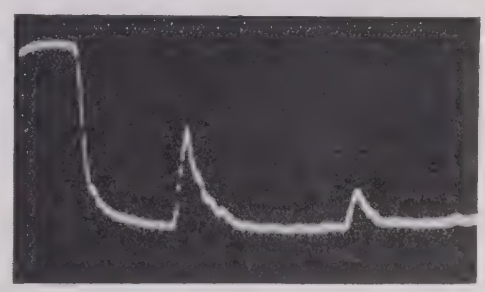


Fig. 6—Single reflection, 2250: 07 PST, July 19, 1948, $T = 75$ μs, 5-km markers.

pulse and the reflection. The simplest procedure is to extrapolate to the base line the linear part of the leading edge of the reflection and measure the time delay from the beginning of the sweep. Accordingly, the time delay of the first reflection is found by counting the number of 30-kc timing markers, which is 20.9. This corresponds to a time delay of 698 μs. To find the total time required for transmission over the almost vertical path (horizontal reflecting layer assumed), we add the time required for transmission of the ground pulse from transmitter to receiver. Since this distance is 2.70 km, the correction is 9.0 μs, assuming ground-wave propagation at the velocity of light. Thus the total time delay was 707 μs. Although the equivalent path in this case is triangular, the base of the triangle is short, compared with the height of the reflecting region. Assuming the triangle to be isosceles, the half angle at the apex is in this case $\sin^{-1} 9/707 = 44$ minutes of arc. It is thus clear that the difference between the hypotenuse and the altitude of the triangular path may be neglected. The virtual height of the one-hop reflection is accordingly found by multiplying the velocity of light by one-half the measured time delay, which gives a value of 106 km. Fig. 6 also shows a two-hop reflection; i.e., a pulse that has made two roundtrips between the earth and ionosphere. Its virtual height, including the base line

correction, is 210 km. One-half the height of the two-hop reflection should equal the height of the one-hop reflection. In this case, the difference is 1 km, which is the over-all error of measurement in the case of well-defined reflections such as those of Fig. 6. It should be observed that the reflection points, or areas, for two-hop reflection do not coincide with those for one-hop reflection. Although the separation is only 0.7 km in a horizontally stratified layer, it would be greater if the layer were concave downward. When the region is not horizontally stratified, the propagation characteristics of the two-hop reflection in general will not be the same as those of the one-hop reflection. A tilted ground plane will likewise affect the paths followed by a multihop signal. These factors must be considered in attempting to account for occasional differences in the virtual heights obtained from a one-hop reflection and various multihop reflections.

The rise time of both the one-hop and two-hop reflections is about 30 μs, which is the same as the rise time of the ground pulse, while the shape of the tail of the one-hop pulse closely approximates that of the ground pulse in Fig. 5. The slight fore-shortening of the two-hop pulse is believed to result principally from the nonlinear response of the diode detector at low signal levels.

Virtual heights determined in the manner outlined above have been obtained from several thousand oscillograms taken on various nights from October 13, 1947, to July 1, 1948. In cases of splitting (discussed below) the first pulse was used. The median virtual height for each minute during which data were taken is plotted in Fig. 7. The time is indicated in minutes after the Pacific Standard Time of the beginning of each run. Perhaps the most noticeable feature of the data is the irregular character of the variations in height, which ranges from 84 to 106 km. During some periods the height remained practically constant for as long as ten minutes. For example, on the January 28 run, the height remained at 96 km ± 1 km from 2108 to 2119 PST. At other times, however, the height varied widely. Thus on October 13, 1947, the height varied between 106 and 93 km in a ten-minute interval. The values which occurred most frequently in the entire group of data were 91 and 96 km, while the average height was 95.2 km.

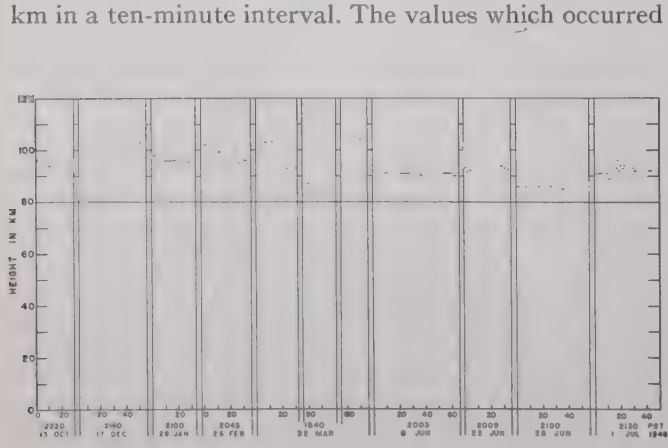


Fig. 7—Nighttime virtual heights at 100 kc.

An interesting variation in virtual height occurred on February 26, 1948. Referring to Fig. 7, one can see that the variation tends to be cyclic with a period of about ten minutes. Possibly it is associated in some way with air currents in the upper atmosphere.

The virtual heights reported here are somewhat higher than those measured by the continuous-wave and lightning-discharge methods described briefly in the introduction. Thus Best, Ratcliffe, and Wilkes[2] found nighttime virtual heights of about 90 km at a frequency of 16 kc using continuous waves over an oblique-incidence path. Others, studying the signals from lightning discharges, found heights ranging from about 60 to 90 km,[3,7] the higher values being obtained at night. The difference may be explained by the fact that lowering the frequency and increasing the angle of incidence reduces the ionization density required for reflection and hence lowers the virtual height in the case of a single layer.

Field Intensity

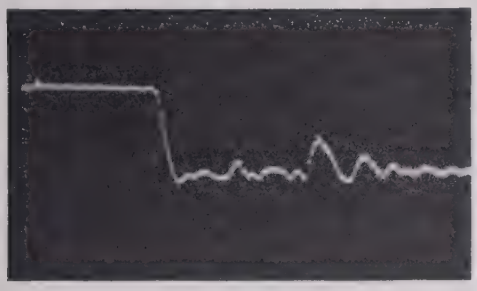
Absolute field intensities were not measured except towards the end of the period, with which this paper is concerned. The field intensities of the one-hop and two-hop reflections appearing on the oscillogram of Fig. 6 were determined by the method outlined in section II and were found to be 33 and 14 μv/m, respectively, at the dipole receiving antenna. The calculated field intensity assuming unity reflection coefficient at the ionosphere is 340 μv/m. If we assume a plane-stratified ionosphere, we can determine the reflection coefficient of the layer by multiplying the ratio of the amplitude of the two-hop signal to the amplitude of the one-hop signal by two and dividing by the magnitude of the ground reflection coefficient. If we take the ground reflection coefficient to be 0.97, the ionosphere reflection coefficient in the above case is found to be 0.88. This reduces the computed field intensity to 300 μv/m, which is about nine times the measured value. A possible explanation of the large discrepancy between measured and calculated field intensities will be given later on in the discussion of the polarization effect.

Echo Splitting

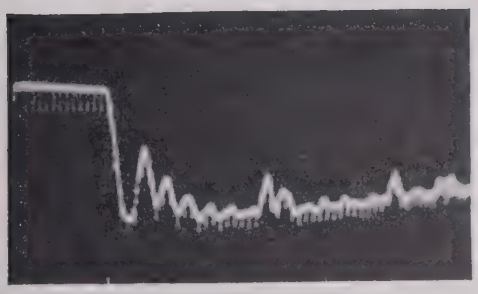
In contrast to the oscillogram of Fig. 6 is the more usual type in which the reflected pulses appear to be split into two or more principal components. An example of the effect is shown in the oscillogram of Fig. 8(a). The first reflected pulse appeared at a virtual height of 91 kilometers. It was followed by a second, weaker pulse at 105 kilometers. A second example of the effect is seen in Fig. 8(b). In this case the pulse time constant was

[7] T. H. Laby, et al., "Reflections of atmospherics by the ionosphere," *Nature*, vol. 142, p. 353; August 21, 1938.

) μs, twice that of the pulse of Fig. 8(a). The first re-tion appeared at 90 km. Its peak was followed by least three additional components at 15-km intervals. two-hop group of reflections appeared at about 178 and shows two distinct peaks separated by about km. A single three-hop reflection appeared at 270 . Further examples of splitting are afforded by the illograms of Figs. 9(a), (b), and (c). In these cases pulse time constant was again 150 μs, but the re-ver bandwidth was roughly seventy per cent of that d to obtain the data of Figs. 8(a) and (b). The timing rker spacing was 10 km, except for the first space ich was 11 km. It will be observed that the rise time the first component in each group was about 45 μs, compared with 30 μs for the previous cases. The chief ference between these oscillograms and those of Figs.) and (b) is the relative lack of separation of the dif-



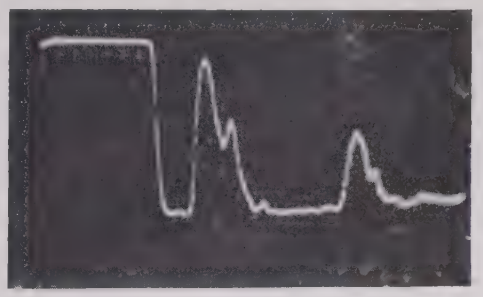
(a)



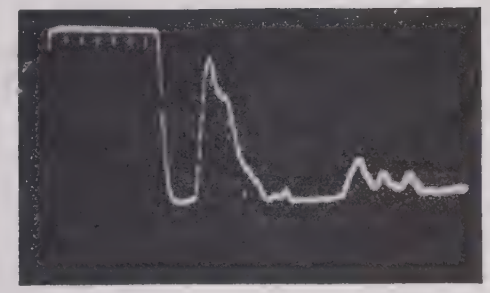
(b)

8(a)—Split reflection, 2106: 49 PST, June 8, 1948, $T=75$ μs, 5-km markers. (b) Split reflection, 2132: 19 PST, June 8, 1948, $T=150$ μs, 5-km markers.

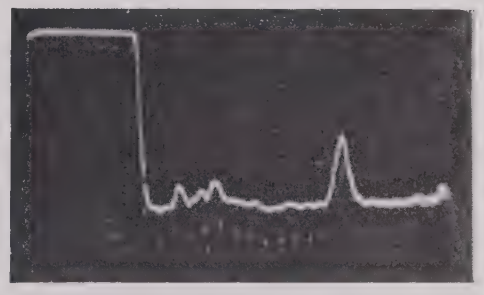
ent components in each group. This overlapping was ompanied by considerable variability in the shape of envelope in the space of a few seconds. Thus, in . 9(a), two distinct peaks appeared on the one-hop ection, and a slight indication of two peaks ap-ared on the two-hop reflection. In Fig. 9(b), three onds later, the second peak in the first group had be-ne only a bulge, but the two-hop group exhibited ee noticeable peaks, the first of which was about one-f the amplitude of its counterpart in Fig. 9(a). A her unusual condition is illustrated in Fig. 9(c), in ich the one-hop group exhibited three relatively ak peaks, while the second group showed only a single ak of normal amplitude. The separations of these peaks varied between about 10 and 15 km on most of the records that have been examined.



(a)



(b)



(c)

Fig. 9(a)—Split reflection, 2155: 22 PST, March 22, 1948, $T=150$ μs, 10-km markers. (b) Split reflection, 2155: 25 PST, March 22, 1948, $T=150$ μs, 10-km markers. (c). Split reflection, 2250: 36 PST, March 22, 1948, $T=150$ μs, 10-km markers.

The variation in the shape of the envelope can be explained by assuming the presence of several component pulses adding together with random relative phases to produce the observed result. Why there should be more than one reflected pulse is more difficult to explain. However, the relative constancy of the separations of most of the multicomponent reflections (of which Fig. 8(b) is a good example) suggests the possibility that the lower part of the E region actually consists of a number of patchy, nonuniform layers, each of which contains "holes" large enough to permit some energy to penetrate (at nearly vertical incidence) to patches of ionization in the next higher layer. Another explanation is that the reflections of longer delay come from patches

of ionization which are at about the same height, but which are not directly overhead. Further experiments aimed at resolving this uncertainty are planned.

Polarization

One of the most interesting and consistent effects observed during the course of these experiments has been the rotation of the plane of polarization of the downcoming pulses, with respect to the plane of polarization of the transmitted pulse. The polarization of the signal transmitted vertically upwards will, of course, be in the direction of the horizontal part of the transmitting antenna, which is oriented fifteen degrees west of true north. At the receiver, located 2.7 km away in a direction 45 degrees west of north from the transmitter, maximum signal strength is obtained with the dipole antenna oriented about ninety degrees from true north. Furthermore, the signal strength with the dipole at right angles to this position is usually zero or a small fraction of the peak intensity, indicating that the polarization is linear, or very nearly so.

One possible explanation of the observed rotation is that the component of signal at right angles to the direction of maximum signal strength is not reflected, or is highly absorbed, while that in the direction of maximum signal strength is reflected with relatively small loss. In order to bring the calculated signal strength into agreement with that measured from Fig. 6, it is necessary for the component of current producing the observed signal to be one-ninth of the total current. Thus the direction of the polarization of the received signal would have to be at an angle of six degrees ($\sin^{-1} 1/9$) with respect to a normal to the transmitting antenna. There are, of course, two such directions separated by twelve degrees. One corresponds to the direction sixty-nine degrees west of south and the other to the direction eight-one degrees west of south.

The variation of field intensity with dipole azimuth was not obtained at the time the oscillogram of Fig. 6 was made. However, approximate measurements of the direction of polarization made at other times indicate that the average direction is east-west with an estimated variation of plus or minus ten degrees. The required direction of eight-one degrees lies within this range, and hence a check is provided on the explanation given above, within the limits of error of measurement, which, in the case of the polarization directions, are appreciable.

In terms of the magneto-ionic theory, one would expect the propagation to be of the "ordinary" component requiring considerably less ionization density for reflection than that required by the "extra-ordinary" component. Even with a relatively sharp ionization gradient, some difference in time delay between the two components would be expected. It would thus be possible for the polarization of the sum of the two reflected components to become more or less elliptical, with the major axis shifted with respect to the direction of t incident plane-polarized wave.

The detailed correlation of the experimental resul with the theory will be postponed until more data a available.

IV. Conclusions

This experimental investigation has shown that t minimum height of the lower region of the nighttin *E* layer is far from constant, varying all the way fro 84 to 106 km during the course of nearly a year intermittent measurements. If low-frequency transmi sion problems are to be solved on the basis of virtu height data, as is done at high frequencies, then mea urements similar to those described in this report mu be made at many different locations and at several fr quencies to determine what variations in height may l expected.

Some evidence has been found to indicate that fr quently at night the region of reflection of 100 kc wav consists of clouds or patches of ionization, rather tha the more nearly uniform ionization characteristic of t regular layers at higher frequencies. If this be the cas the relation between oblique and vertical-incidenc transmission (sometimes called the "equivalence the rem"), which holds for a horizontally stratified laye cannot necessarily be expected to hold at low fr quencies at night. It is therefore highly desirable tha studies be made to determine the nature of the relatio at low frequencies.

The reflection of 100-kc waves is apparently accon panied by changes in polarization. Since the details this effect have not been elucidated, it is importa that further experiments be conducted for this purpos both at vertical incidence and at oblique incidence.

V. Acknowledgments

The author wishes to acknowledge the co-operatio and financial support of the Central Radio Propagatio Laboratory of the National Bureau of Standards carrying out these measurements.

The author is grateful for the generous aid receive during the course of the research from members of t staff of the department of electrical engineering, Sta ford University. Particular mention s ould be made Joseph S. Carroll and William G. Hoover, whose help the design and construction of the high-voltage equi ment was invaluable. The author is likewise indebted Frederick E. Terman, Dean of the School of Enginee ing; Hugh H. Skilling, Executive Head of the depar ment of electrical engineering; and Karl Spangenber for their helpful suggestions. Thanks are due to La rence A. Manning and Oswald G. Villard, Jr., for the many stimulating comments. Mention should also made of the enthusiastic participation of the sever members of the project staff who were engaged in va ous phases of the work.

The Demodulation of a Frequency-Modulated Carrier and Random Noise by a Discriminator*

NELSON M. BLACHMAN†

WITH THE AID of an idealized representation of the discriminator, Middleton[1,2] has shown that a equency-modulated carrier remains intelgible for smaller carrier strengths when arrow-band frequency modulation is used d the limiter is omitted than with a niter and/or wide-band frequency modution. It is, therefore, of some interest to eat the demodulation process as it *actually* curs, in the absence of a limiter. Thus, the scriminator has been taken to consist of vo selective circuits, both fed by the out-ut of the intermediate-frequency amplifier, but peaked at different frequencies, feeding rectifiers whose outputs are subtracted. Each half of the device is treated in the manner used by Rice[3-5] to determine the result of passing random noise and a sine wave through a rectifier; however, there is a correlation between the noise voltages fed to the two rectifiers.

The signal output and the spectral distribution of the noise output are obtained first for quadratic rectification, then in the general case, which is specialized then to linear rectification. The results are applied to a case of rectangular if noise spectrum, and the signal-to-noise ratio is determined for the cases of narrow-band and wide-band FM.

These results are found to be very much like those for the idealized representation of the discriminator;[6] all are compared, along with amplitude modulation,[7-10] in Table I. The optimum signal-to-noise ratio for FM without a limiter is found to obtain with narrow-band FM when the discriminator is designed for no wider a band than necessary; this optimum signal-to-noise ratio differs very little from that for amplitude modulation.

TABLE I

COMPARISON OF MEAN-SQUARE SIGNAL-TO-NOISE RATIOS FOR FM WITHOUT A LIMITER AND AM

System	Demodulator	Weak Carrier ($x \ll 1$)		Strong Carrier ($x \gg 1$)	
		Narrow Band ($f_a \gg F$)	Wide Band ($f_a \ll F$)	Narrow Band ($f_a \gg F$)	Wide Band ($f_a \ll F$)
FM	Quadratic Discriminator	R^2x^2	$\frac{1}{2}(F/f_a)R^2x^2$	$\frac{3}{2}Rx$	$\frac{1}{2}(F/f_a)Rx$
	Linear Discriminator	$2R^2x^2$	$(F/f_a)R^2x^2$	Rx	$2(F/f_a)Rx$
	Idealized Discriminator	$\frac{1}{6}\pi x^2$	—	x	$2(F/f_a)x$
AM	Quadratic Detector	$2x^2$	$2(F/f_a)x^2$	$\frac{3}{2}x$	$\frac{3}{2}(F/f_a)x$
	Linear Detector	$1.83x^2$	$1.83(F/f_a)x^2$	x	$(F/f_a)x$

$2F$ = if bandwidth, f_a = audio upper cutoff frequency, x = carrier-to-noise ratio $\propto 1/F$, deviation ratio $= 3 - \frac{1}{2}F/f_a$ for FM, modulation index = 100 per cent for AM, $1 < R < 4/3$.

* Decimal classification: R148×R361.217. Original manuscript received by the Institute, October 28, 1949; abstract received, March 9, 1949. This abstract summarizes Chapters IV and V of N. M. Blachman, Cruft Laboratory Technical Report No. 31, March 5, 1948.

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† Formerly, Cruft Laboratory, Harvard University, Cambridge, Mass.,; now, Brookhaven National Laboratory, Upton, L. I., N. Y.

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[5] S. O. Rice, "Statistical properties of a sine wave plus random noise," *Bell Sys. Tech. Jour.*, vol. 27, pp. 109–158; January, 1948.

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Discussion on

"Application of Velocity-Modulation Tubes for Reception at U.H.F. and S.H.F."*

M. J. O. STRUTT AND A. VAN DER ZIEL

E. Barlow:[1] Strutt and van der Ziel have given an expression for the minimum noise figure obtainable with a particular adjustment of a three-resonator klystron amplifier. They have, using their notation,

$$N_{\min} = \frac{eI_0F^2}{2kT} R_s\left[\frac{S_{23}^2}{S_{13}^2 + S_{12}^2S_{23}^2R_S^2}\right]. \qquad (14)$$

With the assumptions used by Strutt and van der Ziel, this equation should read, instead,

$$N_{\min} = \frac{eI_0F^2}{2kT} \frac{R_s(S_{12} - S_{13})^2}{S_{13}^2 + S_{12}^2S_{23}^2R_s^2}.$$

The two equations give the same result if the transconductances are calculated from simple bunching theory (a condition not implicit in the derivation of (14)) since, then,

$$S_{13} = S_{12} + S_{23},$$

but if any space-charge debunching is taken into account they are no longer equivalent. This opens up the possibility of getting a zero noise figure (as far as shot noise is concerned). Suppose simple debunching theory be considered, so that the transconductances take the form

$$S_{12} = \frac{1}{2} \frac{I_0}{V_0} \frac{\omega}{v_0} \sin \frac{(hx_{12})}{h}, \text{ etc.},$$

where h is the debunching wave number. It is now no longer true that

$$S_{13} = S_{12} + S_{23},$$

and if, for example,

$$\sin (hx_{13}) = \sin (hx_{12})$$

and

$$x_{12} \neq x_{13},$$

a zero noise figure is indicated by the correct form of (14). For example, if

$$hx_{13} = 2\pi/3,$$
$$hx_{12} = \pi/3,$$
$$hx_{23} = \pi/3,$$

the conditions are satisfied.

A fundamental assumption of Strutt and van de Ziel's paper is that there is complete coherence of the noise fluctuations along the electron beam. It is of interest to investigate how closely this assumption is met in practice. In the example given by the authors, an improvement of 100/1 was obtained by noise cancellation. In order to achieve this performance a "coherence" of 99 per cent is required, and, of course, even more stringent requirements are imposed if better noise cancellation than this is desired. To formulate the coherence requirements more exactly, a correlation coefficient can be introduced. If the noise currents at two points along the electron beam are designated as i_1 and i_2 such that

$$\overline{i_1} = \overline{i_2} = 0$$

and

$$\overline{i_1^2} = \overline{i_2^2},$$

then

$$\overline{(i_1 - i_2)^2} = \overline{2i^2}(1 - \rho_{12})$$

where ρ_{12} is the correlation coefficient,

$$\rho_{12} = \overline{i_1i_2}/\overline{i^2}.$$

This means that, for 99 per cent cancellation of noise currents,

$$\frac{\overline{(i_1 - i_2)^2}}{\overline{i^2}} = 2(1 - \rho_{12}) = 0.01,$$

and a value of $\rho_{12} = 0.995$ is required for the author's example.

Some of the factors which tend to lower the noise correlation in klystron amplifiers are the following:

1. Space-charge debunching
2. Variation of beam coupling coefficient across the electron beam
3. Partition noise
4. The thermal velocity spread of the electrons.

Debunching of the electron beam due to space-charge forces occurs to some extent in all klystron designs, and should be taken into account. If this effect varies across the electron beam, the current-density fluctuations will lose coherence, since those adjustments of the klystron gain, resonator impedance, etc., necessary to cancel fluctuations on the edges of the electron beam will not cancel those near the center, and vice versa. The debunching of the beam depends on beam current and voltage, beam diameter and drift-tube diameter, and on the distance traversed down the beam axis. If the

* M. J. O. Strutt and A. van der Ziel, "Application of velocity-modulation tubes for reception at u.h.f. and s.h.f.," PROC. I.R.E., vol. 36, pp. 19–24; January, 1948.

[1] Formerly, Sperry Gyroscope Company, Great Neck, L. I., N. Y.; now, Rand Corporation, Santa Monica, Calif.

n is of large section relative to $\lambda v/c$ (the "electron elength) the debunching near the center of the n can be characterized by

$$\overline{i_2^2} = \overline{i_1^2}(\cos hx_{12})^2$$

re i_1^2 is the mean square noise current at the en- ce to the klystron drift tube, $\overline{i_2^2}$ is the mean square e current at a distance x_{12} down the electron beam, h is the "debunching wave number" given by

$$h = \frac{173 I_0^{1/2}}{a V_0^{3/4}}$$

re

a = beam radius.

he edge of the beam the image charges along the tube wall almost neutralize the debunching forces, nat, in this case, roughly

$$\overline{i_2^2} \cong \overline{i_1^2};$$

the correlation between the total noise currents at ions 1 and 2 can be roughly estimated as

$$\rho_{12} \cong 1 - \left(\frac{1 - \cos hx_{12}}{4}\right)^2.$$

the amplifier cited by the authors, we can take

$$x_{12} = 10 \text{ cm}$$

$$h = \frac{173\sqrt{0.01}}{a(1000)^{3/4}} = \frac{1}{10a} \ (a \text{ in cm}).$$

asonable value for a would be 0.5 cm. This yields lue of

$$\rho_{12} = 1 - \left(\frac{1 - (\cos 2)}{4}\right) \cong 0.88.$$

, as was pointed out, a correlation coefficient of not han 0.995 was needed to secure the performance red by the authors, the introduction of debunching y brings the efficacy of the proposed klystron design question. Of course, the preceding analysis of the s of debunching is approximate since the de- hing in the center of the beam is somewhat less that indicated, while there will be some debunch- long the edge of the beam, especially if the beam eter is much less than the drift-tube diameter. Of e, the beam must not be made too small or the tube too large, or the debunching will be excessive, eam coupling coefficients to the resonators will be and the effective resonator shunt resistances will w. This points up the conclusions to a study made w-noise-klystron amplifier design at the Sperry scope Company. Almost anything done to im- the coherence of the noise along the beam acts ver the klystron gain or increase the noise figure e compensation, so that very little performance increase is obtained. A study has been made of debunching, and rigorous results including the effects of finite beam diameter and the presence of the drift tube wall have been reached by E. Feenberg. From these results the lack of coherence due to debunching can be estimated, and it turns out to be less serious than the previous rough estimate, but sufficient to make noise cancellation schemes such as that proposed by Strutt and van der Ziel of questionable use.

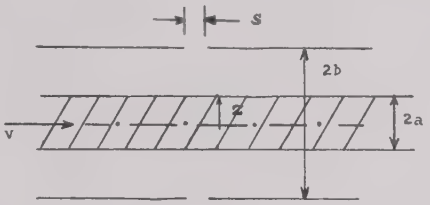


Fig. 1

An analogous effect which must be taken into account is the variation of the beam coupling coefficient across the electron beam. If gridless resonators are used, this coefficient varies as shown in Fig. 1.

$$\mu(z) = \frac{\sin \dfrac{\omega s}{2v}}{\dfrac{\omega s}{2v}} \left[\frac{J_0(jk'z)}{J_0(jk'b)}\right]$$

where

$$k' = \frac{2\pi c}{\lambda v}.$$

If a differential current element dI is considered and the equations of the klystron amplifier applied, a relation will be found for the minimum noise figure, and in that relation will appear

$$\mu_1(z_1), \qquad \mu_2(z_2)$$

and

$$\mu_3(z_3),$$

the beam coupling coefficients the current element experiences in traversing the three gaps. In order to be able to cancel the noise fluctuations of all these filaments with any one klystron adjustment, it is necessary that the ratios

$$\frac{\mu_1(z_1)}{\mu_2(z_2)} \quad \text{and} \quad \frac{\mu_1(z_1)}{\mu_3(z_3)}$$

be the same for all filaments within roughly 2 per cent in the example cited. This puts very stringent requirements on the electron beam and the resonator-gap design. Transverse thermal velocities have been estimated to be sufficient to invalidate this requirement for a typical tube. Any irregularities in the electron-gun

structure or space-charge beam spreading, particularly with nonuniform current distribution across the electron beam, will be deleterious. If fine grids are used, the beam coupling coefficient will be more nearly uniform than with gridless gaps, but will still vary enough to cause trouble and will introduce unwanted partition noise.

There are two distinct effects which are associated with the interception of part of the beam current by the klystron grids.

First, there is "partition noise" in the sense that, if the original beam had shot noise which was reduced by space-charge smoothing, the introduction of a grid intercepting current would partially nullify this smoothing, so that the rms fluctuations after the grid would be greater than those before it. This additional noise would be called partition noise. If the original beam has full shot effect, the partition noise in this sense is negligible.

Second, even though the rms noise output is not increased by the addition of a grid, there will be an element of noncoherence introduced between the noise fluctuations before and after the grid if current is intercepted by it. It is this latter effect which is important in studying the low-noise-klystron problem. Consider a three-resonator klystron using grids as shown in Fig.2. Any current interception on grids 2, 3, 4, or 5 or on the

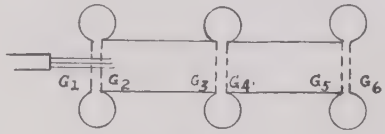


Fig. 2

drift-tube walls will reduce the coherence between the noise currents in the first and third resonators. Assuming full shot-effect fluctuations on the electron beam, the total intercepted current must be $\frac{1}{2}$ per cent or less to keep ρ_{12} above 0.995. Even if there is no current intercepted by the drift-tube walls, the interception per grid in the above example must be less than $\frac{1}{8}$ per cent, while the best figures obtained for fine grids are more like 2 or 3 per cent. This would yield a value of ρ_{12} of roughly 0.90. The authors' example was a klystron having a noise figure, before compensation, of 4,000. With the above figures for grid interception, this could not be reduced below about 400 or 26 db. Strutt and van der Ziel mention that intercepted noise currents affect their conclusions, but state that "if F^2 is near unity . . . no appreciable increase of mean square output current fluctuations will result from these partition fluctuations." This is not the important point; it is the lack of coherence introduced by partition currents which must be considered, as has been pointed out above. The authors further state that "special methods of compensation may be applied, reducing these partition fluctuations to a relatively negligible fraction of over- mean-square output fluctuations." When these spec methods are investigated, it turns out that there is so question as to whether they can be applied to veloci modulation tubes. They are concerned with eliminati of the wrong effect of "partition noise." Consider t case shown in Fig. 3. Let the dc cathode current

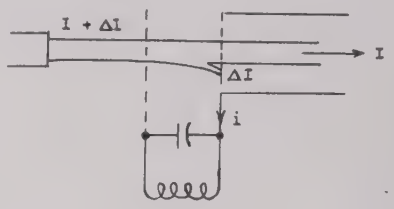


Fig. 3

$I+\Delta I$, and let the current I pass through the gri while the component ΔI is intercepted by grid No. If the noise current i represented only the noise fluct tions of the intercepted current ΔI, it might perh be of some use for compensation, but of course whole idea of velocity-modulation tubes at microw frequencies is that the current I is closely coupled the tuned circuit and i represents the fluctuations both I and ΔI, and, hence, the noise fluctuations in cannot be isolated. If this were not so, the beam co ling coefficient would be identically zero, and grids N 1 and 2 might as well be removed. A further poin that, even if the noise fluctuations in ΔI could be i lated, there is a question as to where a compensat voltage should be applied. If the compensating volt is applied to grid No. 1, a density modulation resu but that in turn introduces an additional noise curr present in both resonator No. 1 and resonator No so that the coherence has not been improved. If compensating voltage is applied to grid No. 2, a vel ity modulation is produced, and one is then in the p tion of trying to compensate a density fluctuation w a velocity fluctuation.

The thermal distribution of forward velocity of electrons leaving the cathode will cause some loss correlation, as pointed out by the authors. This ef can be evaluated in terms of the correlation coefficie Assuming full shot effect, and neglecting space-cha forces in the drift tube, the correlation coefficien given by

$$\rho_{12} = \frac{1}{\sqrt{1+\left(\frac{1}{2}k'x_{12}\frac{kT_c}{eV_0}\right)^2}}.$$

This expression includes both the effects of thermal bunching of density fluctuations and bunching of ini velocity fluctuations. For correlations near unity, can be approximated by

$$\rho_{12} \cong 1 - \frac{1}{2}\left(\frac{1}{2}k'x_{12}\frac{kT_c}{eV_0}\right)^2,$$

d for

$$k'x_{12} = 100,$$

$$\frac{kT_c}{eV_0} = 10^{-4},$$

$$\rho_{12} = 1 - 1.2 \times 10^{-5}.$$

or these values, the thermal velocity spread causes gligible effect. This agrees with the discussion of rutt and van der Ziel on this point. If a klystron amifier is designed for low-noise performance by lowerg the beam voltage and increasing the drift-tube ngth, however, it is quite possible that the thermal locity spread will have to be taken into account. In e such design, a value of $\rho_{12}=0.95$ was found.

In conclusion, it is felt that any scheme such as that ggested by the authors employing noise cancellation mands a careful investigation of noise coherence, and at in particular limitations imposed by (1) spacearge debunching, (2) variations of beam coupling efficient, and (3) partition noise must be carefully instigated. The experimental investigation of noise ncellation made by the writer and referred to by rutt and van der Ziel showed correlation coefficients nsiderably below unity (0.6 to 0.9), and it is felt that is result is in line with the foregoing discussion, since ace-charge debunching and partition noise both isted in the klystron studied.

ax J. Strutt[2] and A. van der Ziel[3]: Mr. Barlow[1] has scussed our paper on noise figures of the three-resonor klystron, and has given an interesting contribution the question of whether the current fluctuations along e beam are completely coherent.

In our paper we showed that reasonably low noise ures could be obtained with the three-resonator klyon, if the following assumptions are valid: (1) simple nching theory can be applied, and (2) complete co-ence exists for the current fluctuations along the am. The second condition was stated explicitly, and ollows from our calculations that the first condition s introduced implicitly. We shall now show that, if first condition is invalid, the second condition is lated also.

Simple bunching theory is not sufficient, if spacearge debunching, transit-time debunching, or variaon of beam coupling along the beam must be taken o account. As shown by Barlow, and as shown for nsit-time debunching in our paper, the fluctuations ng the beam are not completely coherent in these es, so that the second assumption also does not hold. our equations are valid only if the second condition atisfied, it is useless to correct our equation (14) as s done by Barlow. For, if simple bunching theory can be applied, the corrected equation is equal to our equation, whereas both equations do not hold if the first condition, and hence the second one, are not satisfied; therefore, the conclusion of zero noise figure mentioned by Barlow is of doubtful value.

In one of our examples we derived a theoretical improvement of the noise figure by a factor of 100, and although Mr. Barlow does not deny that considerable improvement can be obtained, he has considerable doubt whether an improvement by a factor of 100 is possible. Although we consider Mr. Barlow's discussion very valuable, it still does not give a sufficiently clear indication, of what will be the lowest limit that can be obtained for the noise figure.

In general, we agree with the way in which Mr. Barlow discusses the influence of space-charge debunching and of variation of beam coupling. We think, however, that he overestimates the influence of partition noise. Mr. Barlow mentions that the partition noise due to the current flowing to the grids G_2, G_3, G_4, and G_5 (and to the wall of the drift tube) has to be taken into account. We shall now show, that only the currents to the grids G_2 and G_3 contribute appreciably to the noise figure.

In order to do so, we turn to equation (19) of our paper. After (13) and (16), this holds under the conditions that

$$(S_{23}R_s)^2 \gg 1; \qquad C_2 = 0.$$

Under these conditions, we have, for C_1,

$$\frac{1}{\omega C_1} = \frac{1}{S_{12}}; \quad \text{or, since} \quad Z_1 = (j\omega C_1)^{-1},$$

$$1 - jS_{12}Z_1 = 0.$$

We obtain (19) also if we introduce this relation into (10). For, bearing in mind that $S_{13}=S_{12}+S_{23}$, we have, after (10) and (16),

$$TN_{\min} = N_0 \left| \frac{1 - \frac{S_{13}}{S_{12}}}{jS_{23}Z_2} \right|^2 = N_0 \frac{1}{S_{12}^2 R_s^2},$$

which is the same as equation (19).

This gives the following simple picture of our noise-reducing circuit: The noise figure of a normal klystron amplifier is due to the noise voltage induced in the input circuit by the beam current. In order to improve the noise factor, this voltage must be reduced. This is done with our precircuit, and, after (5), the noise voltage across the input circuit is just equal to zero if the condition $(1-jS_{12}Z_1)=0$ is satisfied. If $S_{12}\neq S_{23}$, the condition for minimum noise is slightly different from the condition of zero noise voltage across the input circuit, but at any rate the major part of the noise reduction is still obtained.

Therefore, it is not very important if some current flows to the grids G_1, G_4, G_5, or G_6, because that hardly

[1] Swiss Federal Institute of Technology, Zurich, Switzerland.
[2] Physics Department, University of British Columbia, Vanver, B.C., Canada.

increases the noise voltage across the input circuit, especially if the beam current is saturated. The currents flowing to the electrodes G_2 and G_3 are the only ones which have to be taken into account, because these currents give rise to partition noise which is uncorrelated to the noise voltage induced in the precircuit, and hence this partition noise will induce a noise voltage in the input circuit, which cannot be compensated by the precircuit. The best solution would be to cut down the currents to the electrodes 2 and 3 as much as possible.

In discussing partition noise in velocity-modulated tubes at microwave frequencies, two facts have to be borne in mind:

1. The beam current is closely coupled to the tuned circuits.
2. The beam current only contains partition noise due to the current flowing to a grid, *after* the beam has passed that grid.

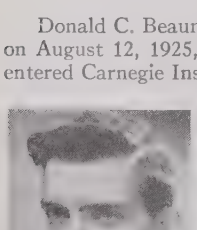
If, for example, I_1 is the beam current before grid G_2 is passed and I_2 the beam current after G_2 is passed, then the noise current flowing through the circuit between grids 1 and 2 does not contain the partition noise generated by the current to G_2, at any rate not if the circuit is of the cavity type. This point is not made sufficiently clear by Mr. Barlow. It is difficult indeed to design circuit which compensates the partition noise due the currents flowing to G_2 and G_3; the best way is to c down these currents to a low value.

In conclusion, we think it to be still worth while investigate what the lowest possible limit will be for t noise factor of the above circuit, although we admit th Mr. Barlow's measurements give some indication th this lowest limit might still be too high.

But even if this were the case, the circuit might st be applied to the traveling-wave tube. Measuremen given by Köempfner indicate that the noise figure of traveling-wave tube is much lower than for a klystr amplifier. For the traveling-wave tube, the noise will b chiefly due to the noise induced by the electron bea into the beginning of the helix. By using a precircui the noise figure might easily be decreased to such a lo value that it becomes lower than the noise of a cryst mixer. One of the advantages of the traveling-wav tube, the very wide-band amplification, would have t be sacrificed, in that case; but the low noise figure wou still make it valuable. It would seem worth while to t the circuit described in our paper in this application a well.

Contributors to the Proceedings of the I.R.E

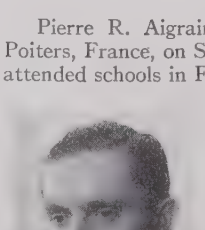
Donald C. Beaumariage (S'46) was born on August 12, 1925, in Bridgeville, Pa. He entered Carnegie Institute of Technology in 1943, and shortly afterwards joined the U. S. Navy, serving until 1946. He attended Cornell University under the Navy College Training Program and received the degree of B.E.E. there in February, 1946. He was employed by the Westinghouse Electric Corporation in the graduate student program, and also did graduate work in the evening school at the University of Pittsburgh during 1946 and 1947. In April, 1947, he became a research assistant at the Carnegie Institute of Technology, and received the M.S. degree in electrical engineering in 1948. At present, he is an assistant research electrical engineer and graduate student at the Carnegie Institute of Technology.



D. C. Beaumariage

Mr. Beaumariage is a member of Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and the American Institute of Electrical Engineers.

Pierre R. Aigrain (S'46) was born in Poiters, France, on September 28, 1924. He attended schools in France, and entered the French Naval Academy in 1942, graduating in 1944. After a short period of active duty with the F.F.I., he came to the United States to receive further flight training at various Naval Air Stations.

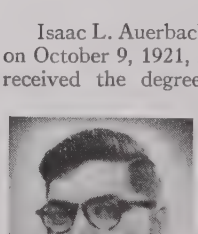


Pierre R. Aigrain

In late 1945 Lieutenant Aigrain went on detached service as a graduate student in the electrical engineering department at the Carnegie Institute of Technology, from which he received the degree of master of science in 1946, and doctor of science in electrical engineering in June, 1948.

Since that time, Lieutenant. Aigrain has been engaged in fundamental research on projects maintained by the French Navy at the Ecole Normale Superieux, in Paris.

Isaac L. Auerbach (S'46–M'49) was bo on October 9, 1921, in Philadelphia, Pa. H received the degree of B.S. in electric engineering from th Drexel Institute Technology in 194 and the degree M.S. in applied phy ics from Harva University in 194 After completing th United States Nav Midshipmen Schoo he attended the Ha vard University–Ma sachusetts Institute Technology Rad School, and then served with the Atlant Fleet as Staff Radar Officer. He was lat transferred to the Combined Researc Group of the Naval Research Laborator where he designed and developed displa and interconnection equipment for th Mark V IFF/UNB Project.



Isaac L. Auerbach

In 1947, Mr. Auerbach joined the Ele tronic Control Company, which later b came the Eckert-Mauchly Computer Cor His work was concerned with acoustic dela memory devices, and he was in charge of th construction of the electronic compute

Contributors to the Proceedings of the I.R.E.

AC. He is now associated with the arch division of the Burroughs Adding chine Company, where he is continuing work on electronic digital computers.
Mr. Auerbach is a member of Eta pa Nu, the American Institute of Elec- l Engineers, and the Association for puting Machinery.

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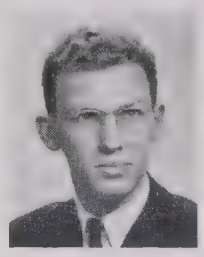
. Presper Eckert, Jr., was born on April 19, in Philadelphia, Pa. In 1941 he re- ed the B.S. degree from the Moore School of electrical engineering of the University of Pennsylvania. During the next year, he taught ESMDT (Engineering, Science, and Management Defense Training) students at the Moore School, and wrote the laboratory experiments for the electronics courses taught to students. During this time he in- ed and had patented a supersonic light e for sound film recording.

P. Eckert, Jr.

ollowing this, he was engaged with P. rumaer in the design and construction device for measuring the concentration aphthalene vapor by means of ultra- t light, for the Barrett Division of the d Chemical and Dye Company. During time, Mr. Eckert was also working John C. Sims, Jr., on the development uipment designed to measure fatigue s in metals for the Southwark Division ie Baldwin Locomotive Works. With completion of his work for Baldwin's, Eckert undertook the designing and ruction of a device for measuring the gth of very small magnetic fields. He also engaged in the solving of a number dar problems, and worked on various es for measuring targets to an accuracy vo yards out of a hundred thousand s. He spent some time on problems in- ng the Differential Analyzer at the re School, and aided in the improvement rque amplifiers (servomechanisms) and e followers for the Analyzer.
n June, 1943, Mr. Eckert received the E.E. degree from the Moore School, which he became chief engineer for NIAC project. In 1945 Mr. Eckert held same position for the design and con- tion of the EDVAC, the successor to NIAC. The following fall, Mr. Eckert Dr. Mauchly formed the Electronic rol Company, now known as the Eckert- chly Computer Corporation.
Ir. Eckert is a member of Sigma Xi. October 19, 1949, he will be awarded the ard H. Potts Medal with Dr. Mauchly ie Franklin Institute of Pennsylvania.

Ralph W. Engstrom was born on October 24, 1914, in Grinnell, Iowa. He received the B.A. degree from St. Olaf, in Northfield, Minn., in 1935, and the M.S. and Ph.D. degrees from Northwestern University in 1937 and 1939, respectively. Dr. Engstrom taught physics at St. Cloud State Teachers College from 1939 until the summer of 1941, when he joined the National Defense Research Council at the University of Michigan. Late in 1941 he joined the Radio Corporation of America, and is at present associated with the RCA Phototube Design Group.



Ralph W. Engstrom

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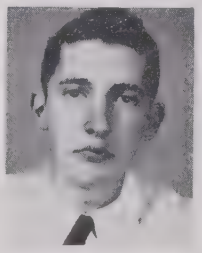
Max Fishman (S'42–A'49) was born in Haverhill, Mass., on April 30, 1921. He received the B.S. degree in 1942 and the M.S. degree in 1943 from Carnegie Institute of Technology, both in electrical engineering. He entered the U. S. Navy in 1944 and served until August, 1946. While in the Navy he was an instructor at the U. S. Naval Academy at Annapolis, after returning from a tour of sea duty on the USS *Gardiners Bay*, AVP-39. Leaving active duty with the rank of lieutenant (junior grade), he returned to Carnegie Institute in September, 1946, and received the D.Sc. in electrical engineering in 1948. Dr. Fishman is a member of Sigma Xi and the American Institute of Electrical Engineers.



Max Fishman

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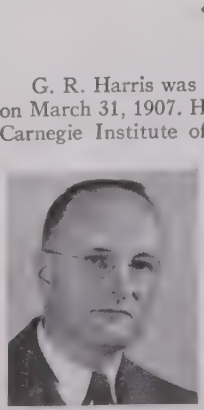
G. R. Harris was born in Pueblo, Colo., on March 31, 1907. He was graduated from Carnegie Institute of Technology in 1927 with the B.S. degree in electrical engineering, after which he was employed by the Bell Telephone Laboratories, Inc. In 1932 he became associated with Jones and Laughlin Steel Corporation, Pittsburgh, Pa., serving in various technical capacities. At present he is a patent attorney for that corporation, having been graduated from Duquesne University Law School in 1941.



G. R. Harris

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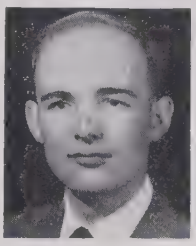
Robert A. Helliwell (S'41–A'45) was born in Red Wing, Minn., on September 2, 1920. He received the degrees of A.B. in 1942 A.M. in 1943, E.E. in 1944, and Ph.D. in 1948, all from Stanford University.



R. A. Helliwell

Since 1942, Dr. Helliwell has been on the staff of the department of electrical engineering of Stanford University where he now holds the position of acting assistant professor. In addition to teaching, he has been engaged in government-sponsored ionosphere research carried out in co-operation with the National Bureau of Standards. During the war he worked on a study of the correlation of direction-finder errors with ionospheric conditions.

Dr. Helliwell is a member of Phi Beta Kappa, Sigma Xi, Tau Beta Pi, and Commission 3 of the URSI.

❖

Arnold R. Moore was born in New York, N. Y., on January 14, 1923. He received the B.S. degree in chemistry in 1942 from the Polytechnic Institute of Brooklyn. From 1942 to 1945, and again during the summer of 1941, he was a research and development engineer at the Lancaster, Pa., Laboratory of the RCA Victor Division, Radio Corporation of America.

Arnold R. Moore

In 1945 Mr. Moore enrolled at Cornell University to work toward the Ph.D. degree in physics. While at Cornell he has been successively a teaching assistant, research assistant, and RCA Fellow in electronics. He is a member of Phi Kappa Phi, Sigma Xi, and the American Physical Society.

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For photographs and biographies of K. Spangenberg, G. Walters, and F. Schott, see page 780 of the July, 1949, issue of the Proceedings of the I.R.E.

Contributors to the Proceedings of the I.R.E

Winston E. Kock (SM'45) was born at Cincinnati, Ohio, on December 5, 1909. He received the E.E. and M.S. degrees at the University of Cincinnati in 1932 and 1933, respectively, followed by the Ph.D. degree in 1934 from the University of Berlin. After a year as Teaching Fellow at the University of Cincinnati, he continued graduate study at The Institute for Advanced Study at Princeton, and at the Indian Institute of Science in Bangalore, India. Following several years as Director of Electronic Research at the Baldwin Piano Company, he joined the Radio Research Department of the Bell Telephone Laboratories at Holmdel, N. J., in 1942, where he first engaged in microwave radar antenna research, and later conducted research on lens antennas for radio relay circuits. In January, 1948, he entered the transmission research department of the Bell Laboratories at Murray Hill, N. J., as research engineer in charge of acoustics.

WINSTON E. KOCK

Dr. Kock is a Fellow of the American Physical Society, a Fellow of the Acoustical Society, and a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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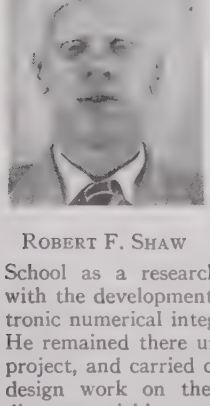
Robert F. Shaw (S'35–A'43–M'45–SM'47) was born at Warsaw, Ohio, on July 10, 1915. He received the A.B. degree in physics from Princeton University in 1937, after which he did graduate work in electrical engineering at the Moore School of the University of Pennsylvania. He taught radio theory for the U.S. Navy in 1942 and 1943, and in the latter year he returned to the Moore School as a research engineer connected with the development of the ENIAC (electronic numerical integrator and computer). He remained there until completion of the project, and carried out preliminary logical design work on the EDVAC (electronic discrete variable computer).



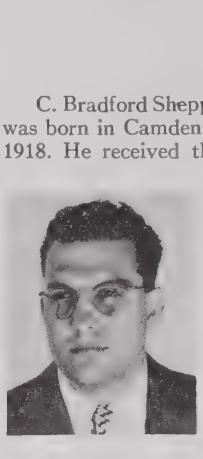
ROBERT F. SHAW

In 1946, Mr. Shaw joined the staff of the Electronic Computer Project at the Institute for Advanced Study, and later that year became a member of the Electronic Control Company, now known as the Eckert-Mauchly Computer Corporation. His work has been concerned mainly with logical design of computers and associated equipment.

Mr. Shaw is a member of the American Association for the Advancement of Science, the Association for Computing Machinery, the Franklin Institute, and Sigma Xi.

❖

C. Bradford Sheppard (S'40–A'44–M'45) was born in Camden, N. J., on October 25, 1918. He received the B.S. degree in electrical engineering in 1941 from the University of Pennsylvania, at which time he was awarded a fellowship at that University. He obtained the master's degree in electrical engineering in 1942, during which year he did part-time teaching in in EMSDT courses, and worked on a Government-sponsored project on antenna field patterns at the University of Pennsylvania. From 1942 to 1946 he was employed by Hazeltine Electronics Corp., first in their antenna design group, and later on test equipment for the Mark V IFF System. In 1946 he returned to the Moore School as a research associate to work on the EDVAC electronic computer.

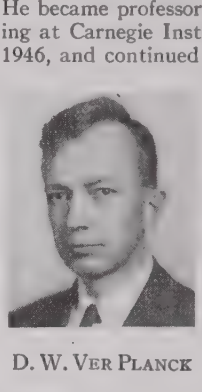


C. B. SHEPPARD

Mr. Sheppard left the University of Pennsylvania in 1947 to join the then newly formed Electronic Control Company, currently known as the Eckert-Mauchly Computer Corp. Here he has worked on acoustic delay and electrostatic storage memory devices, and other electronic computing equipment. He is a member of the American Physical Society, the Association for Computing Machinery, and the American Association for the Advancement of Science.

❖

D. W. Ver Planck was born at Swampscott, Mass., on January 29, 1906. He received the B.S. and M.S. degrees in electrical engineering in 1928 and 1929, respectively, from the Massachusetts Institute of Technology. From 1929 until 1936 he was employed by the General Electric Company. In 1936 he became assistant professor of electrical engineering at Yale University, where he earned the degree of doctor of engineering in 1940, his research being on spark discharges in air. From 1940 until 1942, Dr. Ver Planck was engaged in the development of degaussing of ships at the Naval Ordnance Laboratory in Washington, D. C. From 1942 until 1946 he was officer in the Naval Reserve in charge of gaussing in the Navy Bureau of Ordnan He became professor of electrical engine ing at Carnegie Institute of Technology 1946, and continued in that capacity un until 1947, when was made Geo Tallman Ladd P fessor of Engineeri and head of the partment of mecha ical engineering.

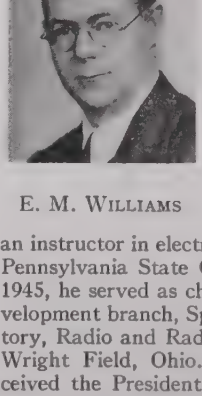


D. W. VER PLANCK

Dr. Ver Planck a member of American Institu of Electrical En neers, the Americ Society of Mecha cal Engineers, American Society for Engineering Educ tion, and Sigma Xi. For his work in Navy he was awarded the Navy Disti guished Civilian Service Award and Navy Commendation Ribbon, and was a made an Honorary Member of the Order British Empire (Military Division).

❖

Everard M. Williams (S'36–A'41–SM' was born in New Haven, Conn., in 1915. received the B.E. degree in 1936 and Ph.D. degree in 19 both from Yale U versity. During summer of 1937 was employed the General Elect Company, and d ing the acade year 1938 and 19 he was the recipie of a Charles A. C fin Fellowship fr this company. Fr 1939 to 1942, he an instructor in electrical engineering at Pennsylvania State College. From 1942 1945, he served as chief engineer of the velopment branch, Special Projects Labo tory, Radio and Radar Subdivision, ATS Wright Field, Ohio. He subsequently ceived the President's Certificate of Me for outstanding service in this post.



E. M. WILLIAMS

In 1945, Dr. Williams was appoint associate professor of electrical engineeri at the Carnegie Institute of Technology, Pittsburgh, Pa., where he is now a prof sor. He is also serving as expert consulta to the Research and Development Board the National Military Establishment.

Dr. Williams was the recipient of 1946 Eta Kappa Nu award. During 194 1948 he served as Chairman of the IR Pittsburgh Section.

Institute News and Radio Notes

Technical Committee Notes

The **Standards Committee** met on June , under the chairman shipof J. G. Brainerd. The following Standards material was acted pon: (a) Additional definitions submitted y the **Electron Tubes and Solid State Devices Committee** were approved; (b) The roposed standards of the **Piezoelectric Crystals Committee** were approved and will e published shortly; (c) Material submitted y the **Radio Receivers Committee** "Tests or Effects of Mistuning and Downward Modulation" was approved; and (d) The tandards Committee approved its subcommittee material on the numbering and indexing of standards. Technical committees and ubcommittees will also be identified by umbers under the proposed system. A subommittee of this committee is studying the uestion of the most economical and thorugh method of circulating IRE Standards o assure the widest distribution. The Standrds Committee is reviewing the scope of wenty-three technical committees. . . . **The ntennas and Wave Guide Committee** held meeting on May 17, Chairman L C. Van tta presiding. Dr. Van Atta recommended he formation of two subcommittees to exedite the work of the main committee, in he preparation of the definitions of terms. . J. Chu was appointed Chairman of the ubcommittee on Antennas, and A. G. Fox as made Chairman of the Subcommittee on Vave Guides. . . . A meeting of the **Electron ubes and Solid State Devices Committee** as held on June 2 at which it was decided to ubmit Power Output High Vacuum Tube Definitions and Small High Vacuum Tube Definitions material to the Standards Comittee for approval. . . . **The Navigation ids Committee,** meeting on June 7, is workng on definitions of Pulse Systems, under Henri Busignies, Chairman. The following asks have been assigned: classification of lectronic methods of navigation systems; resentation of accuracy data; concept of eading time to radio navigation systems; nd coverage of air navigation systems. . . . **he Piezoelectric Crystals Committee** met n May 9. Work on "Proposed Standards n Piezoelectric Crystals" having been comleted this material has been submitted to he Standards Committee for approval. . . n June 8 the **Video Techniques Committee** eld a meeting, at which the following subommittees reported on the status of their ork: Subcommittee V-1, Definitions and ymbols of Video Terms, R. H. Daugherty, r., Chairman; Subcommittee V-2, Utilizaon, Including Video Recording; Methods f Measurement and Test, R. L. Garman. hairman; Subcommittee V-3, Video Sysms and Components, Methods of Measrement and Test, W. J. Poch, Chairman; ubcommittee V-4, Video Signal Transmison, Methods of Measurement and Test, . W. Morrison, Chairman. . . . **Professional roup Chairmen and Sponsors** met on May at Syracuse, N. Y., where reports were iven from three of the six organized Groups, here being no representative present from the Circuit Theory, Audio, or Vehicular and Railroad Radio Communications Groups. Planning reports were presented by sponsors of potential Groups. A progress report on the Quality Control Group was made by Jerome Steen and R. F. Rollman, and Virgil Graham reported for the potential Electron Tube Group. The name of the Broadcast Engineers Group was changed to the IRE Professional Group for Broadcast Transmission Systems, which will include AM, FM, television, facsimile, and the like. O. W. Towner is the Chairman of this Group. . . . On June 7 the **Joint Symposium Planning Group of the IRE/AIEE** held its initial meeting at IRE headquarters to formulate the plans for the 1949 Nucleonics Symposium, tentatively scheduled for October 31, and November 1 and 2, at the Hotel Commodore, in New York City. Harner Selvidge was appointed Chairman of the Joint Symposium Planning Group, representing the IRE. Other members are R. L. Butenhoff, IRE; W. A. Geohegan, AIEE; and G. W. Dunlap, AIEE. Plans were discussed on the basis of a three-day symposium. The tentative program is as follows: First Day, Electronic Aids to Medicine; Second Day, Nucleonics and Biological Tolerances; Third Day, Nucleonics and Industry. Final plans will be announced at a later date.

Newfoundland Added to Montreal Section

Because Newfoundland has become a part of the Dominion of Canada, the territory of Newfoundland and the IRE members residing therein have been added to the Montreal Section of The Institute of Radio Engineers.

Calendar of COMING EVENTS

Summer Seminar, Emporium Section, IRE, Emporium, Pa., August 19–20

AIEE Pacific General Meeting, San Francisco, Calif., August 23–26

1949 IRE West Coast Convention, San Francisco, Calif., August 30–September 2

1949 National Electronics Conference, Chicago, Ill., September 26–28

National Radio Exhibition, Olympia, London, England, September 28 to October 28

AIEE Midwest General Meeting, Cincinnati, Ohio, October 17–21

Radio Fall Meeting, Syracuse, N. Y., October 31, November 1–2

1949 Nucleonics Symposium, New York City, October 31, November 1–2

IRE–URSI Fall Meeting, Washington, D. C., October 31, November 1–2

1950 IRE National Convention, New York, N. Y., March 6–9

Dayton Section Television Lecture Series

To make available recent engineering knowledge of television, the Dayton Section of the IRE last winter conducted a program of television lectures, beginning on January 31, and continuing every Monday until completion, except for the week of the National IRE Convention. The subjects and lecturers for the series were as follows: "Picture Quality," by Madison Cawein, Television Consultant; "Pickup and Transmitting Equipment," by Howard Lepple, Crosley Broadcasting Corp.; "Interconnecting Facilities," by L. W. Morrison, Bell Telephone Laboratories; "Receiving Antenna and Distribution Systems," by Andrew Alford, Consultant; "Receivers," by H. R. Shaw, Colonial Radio Corp.; and "Future Trends," by Joshua Sieger, Freed Radio Corp.

IRE Mentioned in Report on "Scientific Manpower"

The Annual Report on "Scientific Manpower" submitted by the Secretary of the Army to the Secretary of Defense, and covering the period July 1, 1947 to November, 30, 1948, mentions six professional societies as having agreed to participate in a plan whereby their committees will give advice and assistance to the Department of the Army in the fields unique to the societies which the committees represent. Of these six professional societies, The Institute of Radio Engineers is the only one representing electronics or electrical engineering.

IRE to Hold Tenth Annual Seminar at Emporium

The tenth annual summer seminar of the Emporium Section of The Institute of Radio Engineers will be held on August 19 and 20, Roger W. Slinkman, Chairman of the Emporium Section, announced recently. Speakers on Friday, August 19, will include H. G. Clavier, who will discuss carrier power requirements for long-distance communications by microwaves; and Allen A. Barco, whose subject will be problems of television deflection and high voltage supplies. On Saturday, August 20, Michael Landis will speak on modern orthicon camera chain in television; and Gerrard Mountjoy and Henry Appel will discuss design features for a new television receiver.

Following the technical sessions on Saturday morning, the seminar committee will sponsor a picnic at Erskine's Grove to which the public will be invited. George Brunner is Chairman of the Seminar Committee.

Joint Technical Advisory Committee

Summary of Activity—October, 1948 to May, 1949

On October 1, 1948, the Chairman of JTAC, Philip F. Siling, wrote to the Chairman of the FCC, Wayne Coy, offering the assistance of JTAC in obtaining further information and data pertinent to the use of the uhf spectrum for television, or other pressing problems. Mr. Coy's reply, dated October 28, referred to the agenda for the FCC Engineering Conference on VHF Television and FM to be held November 30, 1948, and requested the assistance of JTAC concerning vhf propagation, antennas, co-channel and adjacent-channel interference protection ratios, signal levels required for satisfactory service, and power generation capabilities of transmitters.

In response to this request the JTAC prepared, and offered as evidence at the Engineering Conference, Volume II of the JTAC Proceedings, entitled "Allocation Standards for VHF Television and FM Broadcasting," dated December, 1948. In this report JTAC stated the industry findings on apparatus capabilities but declined to approve or disapprove the FCC T.I.D. reports on vhf propagation, inasmuch as time was not available for the IRE Wave Propagation Committee to review the substantiating data, the methods of using these data, and to give JTAC its expert opinion. The T.I.D. reports were referred by the FCC to the FCC Ad Hoc Committee (on VHF Propagation for Television and FM Broadcasting). JTAC offered, at the Conference, to review the findings of the Ad Hoc Committee when available.

The following Table, taken from JTAC Proceedings Volume II, summarizes the allocation standards recommended by JTAC to the FCC.

Recommended Allocation Standards

Service	Signal Strength	Co-channel protection	1st Adjacent channel protection	2nd Adjacent channel protection
Suburban —Rural TV	500 μv/m (see note)	40 db 25 db*	0 db** 6 db***	— —
Urban TV	5000 μv/m	40 db	0 db**	—
Rural FM	25 μv/m	10 db	−16 db	−40 db
Urban FM	—	10 db	−16 db	−40 db

* Tentative value when co-channel carriers are phase-synchronized.
** Upper adjacent channel.
*** Lower adjacent channel.

In Volume II, the JTAC took note of the technique of phase-synchronization of television picture transmitters, recommending that this technique be taken into account in allocations planning and that further study of it be undertaken. An improvement in co-channel interference of 15 db was mentioned as the best estimate (class-C reliability) then available on the benefit of carrier synchronization.

On December 16, Mr. Siling informed Mr. Coy of the desire of JTAC to assist the commission further in these or other matters. In reply Mr. Coy wrote JTAC under date of December 28, 1948, suggesting the following studies be made:

(1) A study of the number of uhf television channels needed to supplement the vhf channels to provide a sound, competitive national allocation.

(2) A study of the place of color television in the uhf spectrum.

(3) Further study of phase-synchronization of television picture carriers, including possible application to uhf monochrome and color transmissions.

(4) Further study of receiver characteristics, particularly adjacent-channel selectivity, oscillator radiation, choice of a standard intermediate frequency and its effect on channel assignments.

(5) Re-examination of uhf transmitter power-output capabilities.

(6) Preparation of a plan for obtaining more extensive knowledge of propagation on frequencies between 475 and 890 Mc.

On February 18, JTAC replied in a letter addressed to Mr. Coy, expressing the following conclusions:

> JTAC declined to define the number of television stations required in each metropolitan center for national competitive system, on the ground that this was not a question within its province. However, JTAC indicated the range of values which this number might take, based on limitations of channel space within the 475–890-Mc band. The minimum number of stations in each metropolitan center was taken as 2 (since 1 station suffers no competition) and the maximum was taken as 5 (since 6 or more could not be accommodated in the space in the uhf spectrum reserved for television broadcasting). The following table, taken from the letter, shows the number of uhf channels required for a minimum number of stations of 2, 3, 4 and 5 per metropolitan centers, under the condition that the co-channel separation between stations is 150 miles for synchronized and 210 miles for non-synchronized stations. The adjacent-channel separations are 75 miles and 105 miles, respectively.

Allocation Plan No. 1
(Employing phase synchronization on VHF only

2 Stas. Min.	3 Stas. Min.	4 Stas. Min.	5 Stas. Min.
23 (46)	35 (34)	54 (15)	65 (4)

Allocation Plan No. 2
(Employing phase synchronization on VHF and UHI

2 Stas. Min.	3 Stas. Min.	4 Stas. Min.	5 Stas. Min.
18 52)	27 (42)	40 (29)	52 (17)

There are 69 6-Mc channels available be tween 475 and 890 Mc.

The figures in parentheses indicate th number of 6-Mc uhf channels which woul remain for experimentation with other sy: tems of television. JTAC expressed the opir ion that it was not practical to assign vh channels only to primary cities and uhf chan nels only to secondary cities.

Regarding color television, JTAC urge that experimentation with color televisio be actively encouraged, and recommende that adequate space be set aside above 90 Mc and below 6,000 Mc for this purpose. I was recommended that no bandwidth re striction be placed on experimental colo television transmission for the present, unti the relative importance of color and deta in the total value of a television picture ca be determined.

Regarding phase-synchronization of pic ture carriers, the 15-db figure of improve ment was reiterated and its reliability raise from class C to class B.

The opinion was expressed that, in the ory at least, this technique could be applie to uhf transmitters, but the figure of im provement could not be stated pending ac tual tests.

The other items listed in Mr. Coy's lette of December 28 were stated as being unde study. It was recommended that a progran of uhf propagation studies be set up, unde government auspices if possible.

On February 17, 1949, E. W. Allen, Jr. c the FCC wrote to Mr. Siling pointing out a apparent inconsistency in the field strength for satisfactory service cited in JTAC Pro ceedings Volumes I and II. Under date c April 18, JTAC replied, stating that th field strength cited in Volume I were too lo by a factor of 2, since the attenuation due t the vestigial sideband characteristic of th receiver had not been taken into accoun The field strength figures in Volume II wer reaffirmed.

Following the receipt of Mr. Coy's lette of December 28, JTAC circulated a que tionnaire to the IRE and RMA Televisio Systems Committees, concerning the suita bility of the vhf television system standard for use in the uhf service. The replies to th questionnaire were summarized and tran mitted to Mr. Coy under date of April 4. large majority of the replies stated that th vhf standards were applicable witho

ange, while a small number stated that rtain tolerances on carrier frequencies and vels should be examined and changes in- oduced for vhf as well as uhf transmissions. ie opinion was also expressed by several of e respondents that a portion of the uhf ectrum should be reserved for experimen- tion with color and high-definition mono- rome systems.

Under date of April 4 a letter was sent by AC to Mr. Coy commenting on the possi- le use of frequency modulation for the cture transmissions. Previous experiments th this type of transmission were men- ned, and the difficulties encountered from ltipath transmission were described. AC suggested that any plan to consider ing frequency modulation for picture trans- ssions should be preceded by a compre- nsive field test under normal conditions of ltipath transmission. JTAC questioned ether, in view of the present state of owledge on this subject, the time, effort d expense of such a field test would be stified.

Under date of March 30, the JTAC Sec- tary, L. G. Cumming, transmitted to Mr. y copies of replies to a questionnaire cir- lated by JTAC among manufacturers of nsmitting tubes and equipment, indicat- g the amount of power which might be veloped at various frequencies, at that esent time, and what increases might be pected in the future.

On March 22, Mr. Coy sent a letter to AC raising further questions relating to letter of December 28 and JTAC's reply February 18. The questions included:

(1) The preferable arrangement and po- ion of uhf channels within the 474–890 ter band, i.e. as a contiguous block, or erspersed with other assignments.

(2) The justification for assuming a aller adjacent channel separation for chronized operation than for nonsyn- onized operation.

(3) The effect of terrain on propagation.

(4) The absolute values of signal-to-in- ference ratios which give satisfactory vice, with and without synchronization.

(5) Characteristics of receivers, particu- y adjacent-channel interference ratio, eptable signal-to-rms-noise ratio and its tion to signal-to-interference ratio, and se measurements of commercially availa- receivers.

Under date of May 22, JTAC replied to . Coy, recommending that uhf channels assigned in a single block beginning at lower end of the 475- to 890-Mc band. The culty of providing against adjacent chan- interference was admitted. It was stated t considerable improvement in adjacent nnel selectivity could be achieved in re- ers, at some increase in cost. JTAC of- d to secure the absolute values of al-to-interference ratios requested, but mmended against any delay in proceed- with the allocation while these data were g collected. The off-set system (carriers rated about 10,500 cps) of reducing channel interference was mentioned as ring a degree of improvement somewhat erior to that of phase synchronization, less apparatus complexity. JTAC of- d to keep the FCC advised of progress this system. Noise figures of 12 to 13 db were reaffirmed as typical of properly adjusted receivers of the present date, but an additional tolerance of 3 to 5 db was suggested to take account of production variations. The choice of a standard intermediate frequency, with concomitant arrangement of channels to reduce oscillator radiation and image interference, was endorsed, and it was stated that the RMA Television Receiver Committee was being asked to recommend a standard value.

Professional Groups Given Nominating Privilege

Professional Groups now have the opportunity of nominating the Chairman of the subcommittee of the Papers Procurement Committee which is concerned with that Group's field. They may also nominate interested and competent men in their fields as members of the Papers Review Committee and the Board of Editors, all such nominations being subject to acceptance by the Chairman of the Committee concerned, the Editor, and the Executive Committee.

JTAC Officers Announced

The Boards of Directors of the IRE and RMA have appointed Donald G. Fink to serve as Chairman of the Joint Technical Advisory Committee during the period from July 1, 1949, to June 30, 1950; and J. V. L. Hogan to serve as Vice-Chairman for the same period. The JTAC announced that L. G. Cumming will continue to serve as Secretary.

EJC Plans File of Key Engineering Personnel

To provide source material for a Who's Who in engineering research, development, and other scientific operations for use by the National Military Establishment, the Engineers Joint Council, acting through The American Society of Mechanical Engineers, has accepted the task of providing the Office of Naval Research with the names, addresses, ages, and details of professional and scientific qualifications of 100,000 key engineers in all branches of American engineering.

The source file of key engineering personnel thus obtained will provide a valuable tool for solving a variety of technical personnel problems. Its use will diminish disturbances of the national economy, organization of industry, and the personal welfare of engineers, and provide a means by which national resources of technical personnel can be ascertained. The file will also point up weak spots which should be strengthened by education, training, and other means. As a national asset the body of facts will be available to private industrial, educational, and professional-society planning groups, and for other legitimate purposes.

A four-page questionnaire will soon be mailed to 100,000 engineers holding the grade of member or higher in 18 national professional engineering societies. The returns will be collected by the ASME and turned over to the Office of Naval Research of the National Military Establishment for classification.

The project is the result of a conference held in Washington, D. C. last fall, attended by EJC representatives and many other engineering agencies, at which was discussed the need for a list of 25,000 key engineers working in research, development, and other scientific projects who could be called in on a full or part-time basis to work on the broad scientific programs of the National Military Establishment. The task of collecting personal and professional data fell to the EJC as the largest joint agency of the engineering profession. The Engineers Joint Council points out that this will not be just another questionnaire, but one sent to engineers selected from the upper echelon of the profession. The data sought are not intended for general government use, but will go directly to the engineering agencies of the National Military Establishment. As the questionnaire will provide the key to opportunity to professional and patriotic service, engineers selected to receive it are urged to give it serious attention, and to answer all questions fully. The Institute of Radio Engineers is co-operating in this project, and asks its membership to assist in every possible manner.

Television to be Government Monopoly in Australia

Australia will have television in two years, Prime Minister J. B. Chifley announced recently. Run as a government monopoly, television stations will be built in six Australian capital cities—Brisbane, Sydney, Melbourne, Adelaide, Perth, and Hobart. It was also disclosed that each picture will be produced by 625 closely spaced horizontal lines, and compared with 405 lines used by Britain, and 525 in the United States. It was stated that the use of a greater number of lines should ensure a better image than is now available under either the British or American system.

Press Switching Center Aids Foreign Council Meeting

A new press switching center at New York and special direct cable facilities that have been set up in Paris are working together to provide fast and efficient communications to the government and press representatives attending the Council of Foreign Ministers meeting, it was announced recently. One of the special facilities is an exclusive Western Union office, established adjacent to the American delegation conference room in the Hotel Crillon, which flashes thousands of words to the new press switching center in New York. The stories arrive in New York in the form of perforated tape, and are sent through automatic transmitters over direct circuits to press services and newspapers throughout the United States.

National Radiotelephone Network Formed

The formation of the first national network of independent radiotelephone stations for mobile service to the general public was announced recently in a statement filed with the Federal Communications Commission in behalf of the National Mobile Radio System. The new system will offer a practical and unprecedentedly low-cost means of communication between occupants of automobiles, trucks, buses, and other vehicles with offices or homes hundreds of miles distant. The network also has important potentialities as an auxiliary communications system in event of national emergency. By August, the facilities will be in operation between Boston and Washington, and will gradually be extended to encompass all member stations.

The system differs from the mobile radio service offered by telephone companies in that brief messages or conversation are relayed back and forth through the intermediary of a station operator, who receives them by telephone from one end and relays them by radio to the other, or vice versa. Although the new system started with stations on the Eastern seaboard, others in the South and West have joined or have indicated desire to affiliate. It is anticipated that by 1950 at least 100 stations will be active in the network.

Radiocommunication Institute Established in Argentina

The Instituto de Radiocomunicaciones, established last fall at the National University of Tucuman in Argentina, recently announced the appointment of Daniel Eduardo Frias (M'48) as organizer and director. In September, 1948, the Board of Trustees approved curricula for the training of telecommunication engineers, radio technicians, radio fitters, and radio operators. The following departments have also been formed: ultra-high frequency technique, supersonic and hydrophonic waves, air navigation radio, television, radio broadcasting, applied electronics with tubes used in radio communication, and a measuring laboratory. The schedule of studies for the telecommunications courses were adapted especially for the National University of Tucuman by Dr. Frias.

Early in 1949 various measuring equipment and instruments were purchased, and a full broadcasting station was installed by arrangement of Dr. Horacio Descole, President of the University.

First NAB Program Directors' Clinic Held in Chicago

The first National Association of Broadcasters Program Directors' Clinic was held in Chicago on June 27, 28, and 29, on the downtown campus of Northwestern University. All meetings were devoted exclusively to specific, practical discussions for improving program structures at local stations, and aiding program directors to utilize available services to better advantage. Justin Miller, NAB president, spoke to the representative program delegates, stressing the fact that "programs mean audience and audience means income and economic stability."

Industrial Engineering Notes[1]

FCC Actions

The FCC has adopted new rules permitting **private sale of radio broadcasting stations** without public advertising or competitive bidding. The Commission dropped the regulation it has had since 1945 requiring all proposed transfers of all types of broadcasting facilities to be advertised locally for a 60-day period. Under the new procedure, an application for transfer of a radio station will simply be filed with the FCC so that the legal and financial qualifications of the purchaser may be examined. . . . A **total of 2,792 AM, FM and television broadcasting stations** were on the air at the end of May, the FCC reported. . . . The FCC, in an action possibly affecting the use of uhf frequencies by television broadcasting services, today amended its order of May, 1948, relating to the utilization of frequencies in the 475–890 Mc band for television to include **consideration of the allocation of the 470–400 Mc band to multichannel broad-band common carrier mobile radio operation** in lieu of television broadcasting.

National Bureau of Standards Announces New Developments

The National Bureau of Standards recently announced that magnetic fluids, which were originally used in the NBS electromagnetic fluid clutch, have several unique features that make possible important applications of iron-oil mixtures. Studies of the properties of the mixtures reveal that magnetic fluids may be employed to good advantage in a number of ways, including use as variable electrical resistors.

An electronic gating instrument has been designed and constructed at the National Bureau of Standards for the "accurate determination of the deadtime and recovery characteristics of Geiger-Muller counters." Complete information on the new development is available as Research Paper RP1965 at ten cents per copy from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C.

Preferred Parts List Explained by ASESA

To assist further in promoting the electronic component standardization program, the Armed Services Electro Standards Agency has prepared a "Preferred Parts List" for use as a guide in selecting components commonly used in the development and maintenance of electronic equipment. This list contains the electronic components for which preferred form factors, values, and tolerances have been selected. Consideration has been given to cost and production factors; these are reflected in the list by the fact that components included have been screened for the most reasonable tolerances used in the maximum of engineering application and for those which will not appreciably increase the cost of purchase when bought in large quantities.

[1] The data on which these NOTES are based were selected, by permission, from the Radio Manufacturers Association's "Industry Reports," issues of February 18, 25, and March 4, 11, and 18.

Television News

The FCC recently released the long awaited report on radio propagation effect in the frequency range between 50 and 25 Mc, as compiled by a small group of govern ment and industry engineers since the FCC television engineering conference last De cember.

Documents submitted by the committe included the report and four reference vo umes pertaining to its findings. The mai document (Vol. I) contains the followin committee information and recommenda tions:

A. Evaluation of the Random Variation in Field Intensity from Median Levels Du to Local Terrain and Buildings.

B. Method of Combining the Effects o the Spatial and Time Variations of the De sired Signal and One or More Interferin Signals.

Postponement of the lifting of the pre ent TV "freeze" until late fall to allow tim for a new hearing and other proceedings, i cluding consideration of 6-Mc color on bot vhf and uhf channels, was announced by th Federal Communications Commission.

Radio Set Production Declines as Television Rises

The weekly rate of television receiv production in April was the highest yet a tained by the industry, according to a RMA tabulation showing that RMA mem ber-companies manufactured 166,536 T sets during that month.

Radio receiver production in Ap dropped to new low levels. Radio sets r ported for the month by RMA membe totaled 506,409, of which 37,563 were F and FM-AM types. AM set production 468,906 was the lowest since January, 194 when the industry was reconverting fro military to civilian manufacturing.

For the first time RMA members r ported the number of TV receivers with F reception facilities. These sets total 47,264 or about 28 per cent of the TV recei ers produced.

FM and FM-AM set production in Ap was about 62 per cent under the weekly a erage for the first quarter of 1949.

Canadian Radio News

Approximately 3,750,000 radio receive were in use in Canada at the end of 194 and 575,000 sets were produced during t year. Canadian sales of radio receivers ro in numbers in January, 1949, but declined price. A total of 40,794 units valued $3,308,270 were sold, compared with 39,04 sets valued at $3,720,102 in January, 1948

Sales of radio receiving sets by Canadian nufacturers in February totaled 44,268 ts valued at $3,328,642, compared with 833 sets valued at $3,230,740 in the cor- ponding month of 1948, according to a ort received recently by the U. S. Depart- nt of Commerce.

Canadian imports of sets in February ounted to 1,567 units valued at $81,551, npared with 1,379 units valued at $58,529 January. Exports in February totaled 98 sets valued at $106,143, against 2,124 ts valued at $74,920 in January.

Receiving tube production in February aled 309,728 tubes valued at $163,574, npared with 263,582 units valued at 3,612 in January. Canadian manufactur- imported 54,049 tubes valued at $57,916 February, as against 128,050 tubes valued 85,885 in January. Imports of tube parts e valued at $40,372 in February, against ,986 in January.

dio and Television Abroad

Local assembly of radio receivers was ted in **Greece** during March, 1948, and 00 sets were completed during the year. roduction of radio receivers in **Austria** in 8 totaled 96,437 sets, as compared with 46 in 1947 and 127,472 in 1937 .... the end of 1948 there were 475,000 radio ivers in Hungary, 188,000 of them in dapest .... On December 31, 1948, re were 42 radio broadcasting stations in ama, compared with 26 at the close of 7.

Licensed radio receivers in **Denmark** on rch 31 totaled 1,177,608 and represented ncrease of 4.43 per cent over the 1,127,- licensed sets on the same date in 1948.

In **Brazil** the value of imports of radio , parts, and tubes in 1948 dropped to ,128,000 cruzeires from 454,733,000 cru- es in 1947, according to a report received he U. S. Department of Commerce. U. S. nufactured equipment accounted for 86.5 cent of the total in 1947 and 76.6 per cent 948. Holland supplied 10.5 per cent of the Brazilian radio imports in 1948 and 3.7 per cent in 1947. The United Kingdom accounted for 8.0 per cent of the Brazilian total in 1948 and 4.9 per cent in 1947. .... Radio receiver production in the Bizonal area of **Germany** totaled 407,472 in 1948, according to statistics received by the U. S. Department of Commerce. Production of receiver and amplifier tubes and valves totaled 3,111,885 units in 1948. Production of radio sets in **Austria** totaled 96,437 in 1948, compared with 21,246 in 1947 and 127,472 in 1937. .... Production of radio receivers in **Japan** in 1948 averaged 67,749 sets monthly, according to statistics received recently by the U. S. Department of Commerce. This compares with monthly averages of 64,870 receivers and 43,221 sets in 1947 and 1946, respectively. **Transmitter** production averaged 168 monthly in 1948, compared with 227 in 1947 and 97 in 1946. .... A bill was recently introduced in the legislature of **Cuba** to "regulate industries" including the manufacture and assembly of phonographs, radio and television transmitting and receiving apparatus. The measure would grant for a period of ten years exemption from consular fees, duty, taxes, charges, and surcharges on imports by these industries of parts and pieces used in manufacture or assembly. .... Television is still in an experimental stage in **Sweden** and experts agree it will be from two to three years before programs for the public are introduced, according to a report on television prepared by Valdemars Kreiebergs, of the American Embassy at Stockholm. "Televising of programs to the public will not be introduced before Sweden and a number of other countries of the European Continent, including Denmark, Norway, Holland, Belgium and Switzerland, have agreed on standardization, i.e., the number of lines to a surface unit in the transmitted picture," the report points out. Different types of foreign receivers have been tested but it has not yet been decided which type will be adopted for Sweden. A modified U. S. system, a receiver using 625 lines instead of the 525 lines used in this country, is "seriously considered." Current television research in Sweden is being carried on under the supervision of the Television Research Board. The Board was established in November, 1947, at a conference of representatives of interested Swedish agencies and firms. Several press releases from the Television Board mentioned the building and installation of a television transmitter in the Stockholm Institute of Technology. Its power is 1 kilowatt, and at present it operates with a carrier frequency of 80 Mc. Later one will test operation on frequencies of 60 and 200 Mc.

RMA Elects New Officers

At its "Silver Anniversary" convention held on May 16–19, at the Stevens Hotel, Chicago, the RMA elected and appointed the following officers:

President

R. C. Cosgrove, Exec. Vice Pres., Avco Mfg. Corp., Cincinnati, Ohio (new-4th term)

Executive Vice-President and Secretary

Bond Geddes (A'41), 1317 F Street, N.W., Washington, D. C. (re-elected)

Vice-Presidents

G. M. Gardner, Chairman, Wells-Gardner & Co., Chicago, Ill. (re-elected)

R. E. Carlson, Vice-Pres., Tung-Sol Lamp Works, Inc., Newark, N. J. (re-elected)

W. J. Barkley (M'29–SM'43), Exec. Vice-Pres., Collins Radio Co., Cedar Rapids, Ia. (re-elected)

A. D. Plamondon, Jr. (SM'46), Pres., Indiana Steel Products Co., Chicago, Ill. (re-elected)

A. Liberman (SM'49), Pres., Talk-A-Phone Co., Chicago, Ill. (new)

Treasurer

Leslie F. Muter, The Muter Co., Chicago, Ill. (re-elected)

Director of Engineering Department

W. R. G. Baker (A'19–F'28), Vice-Pres., General Electric Co., Syracuse, N. Y. (reappointed)

ooks

uum Tube Amplifiers, edited by George 'alley, Jr., and Henry Wallman

ublished (1948) by the McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 733 pages page index+xvii pages. 186 figures. 6¼×9¼. 00.

This book is volume eighteen of the nty-eight-volume Radiation Laboratory es instituted to record wartime contribu- s to the art. As one can readily surmise, as been necessary to fill in wartime con- itions with considerable background ma- l, in order to give the volume continu- It is believed that the book would have more useful and valuable, had the back- nd material been carefully annotated footnote references.

The preface states that:

> The amplifiers discussed in this volume are designed to have extreme values in one of several of the pertinent characteristics: bandwidth, sensitivity, linearity, constancy of gain over long periods of time, etc.

This represents a good summary of the material included, to which should be added that the manner in which the material is treated ranges from profound and precise mathematics to handbook material, such as JAN characteristics and tolerances. It is, therefore, difficult to recommend the book to specific groups; rather, it will probably be found that certain parts of the book will be valuable to widely divergent interests. Since there is no other book of comparable scope now available, those actively engaged in amplifier development and design will find it very useful. The material included will doubtlessly also be valuable to college professors in preparing classroom lectures. Frequent tabulations, summaries, and circuit diagrams are of considerable utility.

Chapter Four, "Synchronous and Staggered Single-Tuned High-Frequency Bandpass Amplifiers," Chapter Five, "Double-Tuned Circuits," and Chapter Fourteen, "Measurements of Noise Figure," contain material of sufficient interest and importance to warrant special mention. Dc restorer circuits are treated only briefly. The chapter on dc amplifiers does not include consideration of flicker effect on stability, and a nine-page index hardly seems adequate for a book of this size and scope.

L. J. Giacoletto
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Princeton, N. J.

Cosmic Ray Physics, by D. J. X. Montgomery

Published (1949) by the Princeton University Press, Princeton, N. J., 357 pages+12-page index +viii pages. 123 figures. 6×9. $5.00.

There is perhaps no field of research which has undergone as rapid and revolutionary a development during the past two decades as has the study of cosmic rays. During this short period the original concept that the primary cosmic radiation consisted of photons changed first to an electron hypothesis, then to a positive proton theory (with occasional suggestions of negative protons), and more recently to the proton plus heavier charged particles concept. To write a book on a subject which is undergoing such rapid changes is a hazardous undertaking; the book might well be out of date before it is published.

Since "Cosmic Ray Physics" by D. J. X. Montgomery is the most recent book published on cosmic rays to the reviewer's knowledge (at the time of writing), it is also the most up to date. In general, the author has handled the broad scope of the subject very well by covering both the early work, as well as most of the important recent work (up to the publication date of September, 1948), including a chapter on the very recent findings on charged particles heavier than protons in the cosmic radiation.

The basis for the material in the book is a series of lectures given by Professor Marcel Schein of the University of Chicago at Princeton University in the spring of 1948. Professor Schein is one of the outstanding authorities in the field of cosmic rays, having worked with A. H. Compton at Chicago for many years, and, during the past ten years, having himself made many notable contributions in this field. The original notes prepared from Professor Schein's lectures have been considerably expanded, and much material has been added, including some contributions by Drs. Neils Arley and Shuichi Kusaka, both outstanding theoretical physicists in the field of cosmic rays.

The book is written from the experimental point of view, and only such theory as is necessary to the fundamental concepts and the immediate experiments is included. Several chapters are devoted to the experimental equipment and techniques used in cosmic-ray work. A chapter on the intensity of cosmic radiation discusses the absolute intensity as well as the geomagnetic effect and its relation to the directional intensity. The simple Stoermer theory of the motion of a charged particle in the field of a magnetic dipole (the earth) is given in an Appendix.

Mesons in the cosmic radiation are discussed very completely under the heading of "Hard Component," and the types of meson —their charge, mass, lifetime, spin, etc.— are discussed in great detail. The chapter on the soft components, written by Dr. Arley, reviews the subject of the electron and photon components without recourse to the very successful but somewhat complex cascade theory for which Dr. Arley was partly responsible.

The book is well organized and is profusely illustrated with excellent reproductions of cloud chamber and photographic emulsion tracks, as well as many curves and line drawings. One of the Appendices contains the very useful range-energy and range-momentum curves for protons, mesons, and electrons, which were prepared by Princeton as part of an Office of Naval Research project.

"Cosmic Ray Physics" should prove valuable both to the individual working in other fields who has a good physics background and a casual interest in cosmic rays, as well as a reference text for the worker in the field. It should also serve as an excellent textbook for the study of cosmic rays.

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Radio at Ultra-High Frequencies, Volume II

Published (1948) by the *RCA Review*, RCA Laboratories, Princeton, N. J., 485 pages+x pages. 350 figures. 6×9. $2.50.

This is a well-organized reference work for the engineer engaged in development and research on systems and components in the ultra-high and higher frequency portions of the spectrum. Papers of major importance by RCA authors, which were published in various technical journals during the years 1940 through 1947, and covering in general the frequency range 300–3000 Mc, are here grouped in sections covering antennas and transmission lines, propagation, reception, radio relays, microwaves, measurements components, and navigational aids. A complete bibliography of the ultra-high-frequency field is appended in addition to summaries of all papers appearing in Volume I of "Radio at Ultra High Frequencies."

For those engineers and scientists unfamiliar with ultra-high-frequency techniques, components, and systems, this book provides under one cover material which previously was available only after much laborious library research.

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Sperry Gyroscope Co.
Great Neck, L. I., N. Y.

The Radio Amateur's Handbook

Published (1949) by the American Radio Relay League, West Hartford, Conn., 605 pages+10-page index. 1,651 illustrations. 6¼×9¼. $2.00.

This hardy perennial in the field of technical literature, now in its twenty-sixth edition, remains a mine of practical circuitry for the radio beginner, the active amateur, and the professional engineer alike. In no sense a substitute for a university course in electrical engineering, it nevertheless presents a surprising amount of fundamental radio theory in an easily absorbable form, with a strong emphasis on physical concepts, rather than mathematical derivation and exposition. From the opening chapter reviewing briefly the history, scope, and purpose of the amateur radio movement to the final tabulation of vacuum-tube data, the book hews closely to its avowed purpose of presenting within one cover the basic information needed to enter the field of amateur radio, and to remain abreast of those developments in the field of communications most readily adapted to the amateur service.

Following two chapters devoted to a compressed but lucid exposition of electrical laws and vacuum-tube principles, chapters four to ten are given over to applications principles in practical equipment. Act component values for typical circuits given, together with photographs clarify mechanical layout, and a considera amount of the engineering "know-ho needed in practical design work is conve in this sugar-coated way. The succeed five chapters are devoted to vhf and principles and practice, with considera use of circuitry brought to high developm during the war, such as broadband rf am fiers, triode mixers, etc. Some principles klystrons, magnetrons, waveguides, a cavity resonators are also expounded, the coverage of this field is spotty, and lit practical equipment for microwave work described.

Chapter 16 is concerned with measur equipment, mainly conventional frequer meters, crystal calibrators, grid dip oscil tors, and similar devices. The next se chapters are given over to actual assem and operation of a complete amateur stati and a dissertation on the organization function of the American Radio Re League. Two chapters on miscellane and vacuum-tube data of considerable ref ence value complete the text, and there i fairly comprehensive index. A "catalog section, in which many of the more pro nent component manufacturers list such their products as are well adapted to an teur service, completes the volume.

It is perhaps imprudent to point small deficiencies in a technical publicat with an all-time circulation figure over 2,000,000 mark, yet it may be mentioned passing that treatment of low noise rf inp circuits, double superheterodynes, sup selective if circuits, and single sideband s pressed carrier techniques is rather ina quate, to say the least, and the question television interference, currently the m vexing problem facing the active amate has been merely skimmed. With these f qualifications, the book is an excellent s cessor to previous volumes in the series, a practically a "must" for those participati in the amateur hobby.

L. JEROME STANT
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New York 14, N.

Radio Fundamentals, by Arthur L. Albert

Published (1948) by the McGraw-Hill Book C 330 W. 42 St., New York 18, N. Y. 583 pages+11-p index+vii pages. 320 figures. 6¼×9¼. $4.50.

In his preface, the author indicates th this new book is designed primarily for b ginning students, for radio technicians, a for radio amateurs, rather than for advanc radio engineers. Within the limitatio he prescribed for himself, he has done w although portions of his material may beyond the comprehension of the gro for whom he is writing. Although the imp tant aspects of radio have all been dealt wi in a manner that should be understood most readers, the use of complex algebra those cases where mathematics is used see to be a sophistication incommensurate wi the avowed level of the text.

Chapter One, "Fundamentals of Acou tics," presents an interesting summary architectural acoustics as applied to pro lems of studio design, and also a gener

atment of microphones. The next three apters, entitled "Electrical Fundamens," "Series and Parallel Resonant Groups," l "Power Transference and Impedance atching" respectively, serve as a review of basic electrical theories underlying most dio work. The other chapters deal with nsmission lines, cables, and networks; cuum tubes; rectifiers; voltage and power plifiers; oscillators; modulation and dedulation; radio transmitters and receiv; and antennas and radio transmission. e use of numerous sketches to illustrate operation of the important modulators d demodulators makes the chapter on that ject particularly noteworthy. A large mber of illustrated examples is given throughout the text, serving to indicate how problems that arise in radio are analyzed and solved. In some cases, however, the analytic basis for the calculations has not been included.

A rather serious weakness of the text is the looseness in the discussion of the equivalent plate circuit representation of a vacuum tube, and also the subsequent discussion of the voltage gain of an amplifier. Equation 80 on page 273 gives the gain of a single stage amplifier as positive, thereby ignoring the 180-degree phase shift through the amplifier. All subsequent gain formulas suffer in the same way. Thus one cannot tell from equation 115 whether the output from a cathode-follower is in phase or out of phase with the input. While this may be a minor point when discussing the operation of a single stage, the relative phase is frequently more important than the gain. Several obvious errors have eluded the proofreaders. For example, the photograph on page 113 is upside down. Also, on page 129 it is stated that G, the shunt leakage of a transmission line, is given in ohms per unit length, instead of mhos.

On the whole, however, this book is one of the better books on the beginner's level, and should find an important place in the literature of radio.

SAMUEL SEELY
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RE People

Philip J. Senn (A'43–M'46), an active member of the Dayton Section of the IRE, died in an automobile accident while on his way to attend the Cincinnati Spring Technical Conference on Television.

Born in Chicago, Ill., on June 17, 1916, Mr. Senn studied at the Lewis Institute and also took courses in special military communications and radar. In 1942 he joined the staff of the Signal Corps Aircraft Radio Laboratory at Wright Field, Dayton, Ohio, transferring the following year to the Signal Corps Aircraft Radio Engineering School in Dayton as an instructor. Since April, 1946, he had been associated with the USAF Air Matériel Command, Aircraft Radiation Laboratory, as a radio project engineer on the research and development of guided missile radio control equipment.

Arthur V. Loughren (A'24–M'39–SM'43–4) has been elected vice-president in rge of research of the Hazeltine Elecics Corp.'s Board of Directors. **Orville Dunning** (A'34–SM'44) was made vicesident in charge of engineering; and nes F. Willenbecher (SM'44) was chosen e-president in charge of manufacturing. Mr. Loughren has been on the Hazeltine f since 1936, during which time he has le many contributions to the theory and ctice of television. Previously he was ociated with the General Electric Co. as nsultant on vacuum-tube problems, and ervisor of broadcast receiver develop-t, and later with the RCA Manufactur-Company, where he was in charge of ing and inspection.

Mr. Dunning was chief engineer for the ora Phonograph Co.; then became manager of the Telediphone department of the Thomas A. Edison Co., and, in 1940, chief engineer of the Gray Manufacturing Co. Joining Hazeltine in 1941, he has been responsible for the administrative and technical supervision of the company's engineering work under military contracts.

Before he joined Hazeltine in 1942, Mr. Willenbecher served on the staff of the RCA Manufacturing Co., and also of the Philco Corp. Prior to this advancement, he was manager of the production division.

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J. E. Shepherd (A'36–SM'44–F'48) was recently elected the 1949–1950 president of the Technical Societies Council of New York, Inc. He is the fourth president of the council, which was organized in 1946 and has a membership of seventeen engineering, scientific, and technical societies in the metropolitan area, representing more than 25,000 members. Dr. Shepherd, a director of the IRE, has been active in the New York Section activities, serving as Chairman in 1947–1948, and Secretary in 1944–1945. He was given the Fellow award in 1948 "for his contributions to the development of airborn radar armament and for his active participation and leadership in the functions of the Institute."

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Louis G. Pacent (A'12–M'15–F'47), president and technical director of the Pacent Engineering Corp. in New York, N. Y., has been appointed consulting engineer by Plessey International, Ltd., in Ilford, Essex, England.

Mr Pacent was born on June 23, 1893, in New York, N. Y., and he received the I.E.E. degree from Pratt Institute in 1916. From 1915 to 1917 he was advertising manager for the IRE, and he has also been a member of numerous Institute Committees. In 1947 he received the War Department Certificate of Appreciation for his services to the Signal Corps.

A Fellow of the Society of Motion Picture Engineers, the Radio Club of America, and the AIEE, Mr. Pacent is also a member of the last organization's Board of Examiners, and a member of the Acoustical Society of America and the Engineers Club of New York.

❖

Allen B. DuMont (M'30–F'31), president of the Allen B. DuMont Laboratories, Inc., received the honorary doctor of science degree, conferred by the Brooklyn Polytechnic Institute on June 15, 1949.

Ralph E. Hantzch (M'48), superintendent of manufacturing of exploration and production instruments and equipment at the Shell Oil Co. in Houston, Tex., died early this year.

Mr. Hantzch was born in Milwaukee, Wis., on April 2, 1896. In 1918 he was employed as an instructor at the Signal Corps School of the University of Wisconsin, following which he served as a student assistant in physics and radio for the next two years. From 1920 to 1921 he was an instructor in physics and elementary mathematics at the University of Wisconsin, receiving the B.S. degree in electrical engineering at the end of that period.

Following his graduation, Mr. Hantzch joined the Bell Telephone Laboratories as a designing engineer. He left in 1927 to become chief engineer at the Best Manufacturing Co. In 1930 he resigned to work as a private consulting engineer on theater acoustical apparatus. Six years later he became a development engineer for the T. A. Edison Co. He joined the Shell Co. in 1938.

Samuel Lubkin (SM'46) has been appointed consultant to the machine development laboratory, a division of the National Bureau of Standards' applied mathematics laboratories. There he will advise on the logical, mathematical, and engineering aspects of electronic computers, particularly the development of the NBS Interim Computer and a full-scale computer for one of the Army agencies.

Beginning his career in 1929 as a design engineer and test supervisor for the Otis Elevator Co., Dr. Lubkin left ten years later, and the following year became director of the Philadelphia Signal Depot's inspection laboratory, acting as general consultant on technical matters to the Signal Corps Inspection Agency and the Signal Corps Procurement Agency. In 1946 he was appointed engineer-in-charge of computing machines at the Aberdeen Proving Ground, leaving to become engineer-in-charge of the digital computer section of the Reeves Instrument Corp. the following year. At the same time, he was lent as consultant to the University of Pennsylvania for research on the EDVAC and ENIAC. Dr. Lubkin left the Reeves Instrument Corp. in 1948.

Inventor of a number of control devices and author of many technical papers, Dr. Lubkin has conducted extensive research in electrical, electronic, and mechanical engineering, and applied mathematics. His work has included experimental and mathematical investigations on stresses and vibrations in mechanical structures, analyses of transients in motor-speed control arrangements, programming and design of electronic computers, and instrumentation.

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Wesley Tate Guest (A'32–SM'47), formerly a colonel in the Army Communications Service Division of the Signal Corps, was promoted recently to the rank of brigadier general.

General Guest was born on March 18, 1900, and received the engineering degree from Cornell University in 1921. The following year he attended the Army Signal School and in 1928 and 1929 he did graduate work at Yale University, receiving the M.S. degree. In 1946 he was at the National War College.

Wilson A. Maisel (A'46), head of the electronics department of the Frank H. Parks Co. in Portland, Ore., died early this year.

Mr. Maisel was born on November 5, 1918, in Deming, N. M., and was educated at the Stayton High School in Stayton, Ore., graduating in 1936. After working as a technician in a radio service shop, in 1940 he became assistant to the superintendent of the Instrument Laboratories, Inc., of Seattle, Wash., where he was in direct charge of the design, manufacture, and servicing of marine and industrial electronic instruments.

William Webster Hansen (A'39–F'47), radar pioneer, died recently of a heart attack after a long illness.

Born in Fresno, Calif., in May 1909, Dr. Hansen was graduated from Leland Stanford University at the age of twenty, and he became a faculty member the next year, remaining there until his death. Early in his teaching career he taught several advanced physics courses which he himself had not taken as a student. Later he included them in his studies for the doctorate, which he received in 1932.

In 1937 Dr. Hansen began work on a device to prevent airplanes from flying into mountains, and from this research came the klystron, one of the most important elements of radar. Afterward Dr. Hansen and his co-workers developed other pieces of radar equipment, including the rhumbatron, which is also used in atom-smashers.

In 1944 Dr. Hansen was awarded the IRE's Morris Liebmann Memorial Award for the "application of electromagnetic theory to radiation antennas, resonators, and electron bunching, and for the development of practical equipment and measurement techniques in the microwave field." During the war he worked an eight-hour shift in the Massachusetts Institute of Technology's radiation laboratory; then commuted to the Sperry Gyroscope Co.'s laboratory on Long Island for another eight-hour stretch. It was as a result of this wartime schedule that he became ill. He entered a hospital last fall on his return from attendance at an international atomic physics conference at the University of Birmingham in England.

At the time of his death he was professor of physics at Stanford and director of the Univerity's microwave laboratory. The billion-volt linear accelerator and atom smasher now under construction on the Stanford campus was planned and designed by him. He was a member of the National Academy of Science.

Irving G. Wolff (A'27–F'42), director of the Radio Tube Research Laboratory at the RCA Laboratories, received the Distinguished Public Service Award of the Navy Department, the highest honor bestowed on a civilian by the Navy, in recognition of his achievements in electronics and radar.

Born in July, 1894, in New York City, Dr. Wolff received the B.S. degree in physics from Dartmouth College in 1916 and the Ph.D. from Cornell University in 1923. During the interim he taught at Iowa State College in 1919 and 1920, and at Cornell from 1920 until he received the doctorate. In that year he did research on polarization capacity for the Heckscher Research Council, joining RCA in 1924 as a member of the technical and test department. From 1930 1941 he was with the research division the RCA Manufacturing Co. in Camde N. J., and since then he has been with t Laboratories.

Dr. Wolff's specialization is in proble dealing with microwaves, radar, and avi tion. He has been a member of a number Institute Committees and represented t IRE on the American Standards Assoc tion's Sectional Committee on Acoustic Measurements and Terminology. A fellow the American Association for the Achiev ment of Science and of the Acoustical So ety of America, he also is a member of t American Physical Society.

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Alfred C. Viebranz (A'48), formerly go ernment sales representative for the ele tronics division of Sylvania Electric Pro ucts Inc., at Boston, Mass., has been a pointed their special representative at Was ington, D. C. In addition he will act a technical consultant in all phases of go ernment relations. Prior to joining Sylvan he was in the submarine service as a lieute ant in the U. S. Naval Reserves. During t war he was chief engineer and executi officer on the *U.S.S. Haddo*, and w awarded two silver stars and a bronze st for combat duty in the South Pacific. Lat he served as electronics officer on the *U.S. Sarda*.

A native of New Rochelle, N. Y., M Viebranz received the B.S. degree in physi from St. Lawrence University in 1942, wh serving the United Press as a staff co respondent. Later he attended the postgra uate school of the United States Nav Academy, and was graduated as a commu cations engineer.

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The College of Wooster conferred t honorary degree of doctor of science on V tor J. Andrew (A'25–M'30–SM'43) at seventy-ninth annual convention on Ju 13, 1949. Dr. Andrew was graduat from the College in 1926, and received t Ph.D. degree from the University of Chica in 1932, in recognition of work there in phy ics. He is now chairman of the Board Directors of the Andrew Corp., which established in 1936.

Ambrose C. Kibler (A'44), formerly chief engineer for radio stations WNEX and WNEX-FM, in Macon, Ga., was fatally injured in an automobile accident on May 16, 1949. Born in South Carolina in 1915, he received the A.B. degree from Newberry College in 1934. He became a radio technician by studying in his spare time, and was an instructor at the radio training school maintained by the Signal Corps, WRASC, in Warner Robins, Ga., from 1942 until the end of the war. Mr. Kibler's title was civilian radio training administrator.

William E. Osborne (A'41) has been des- ated chief of the electronic and guidance ision of the Marquardt Aircraft Co. is division has been an operating unit ler Mr. Osborne since 1947, and has just ently been moved to Van Nuys, Calif.

Mr. Osborne has been engaged in elec- nic and radar work since 1925, when he eived the E.E. degree from Queens Col- e, in Melbourne, Australia. Now a U. S. zen, he was a British Navy captain dur- 1939 and 1940, and subsequently served a radar liaison officer to both the United tes and British Governments. In 1945 he s associated with Gilfillan Brothers, Inc., principal radar design engineer until 1947, en he organized the division of which he ow chief.

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J. W. Head (A'44–M'48) was awarded honorary LL.D. degree by Piedmont llege, Demorest, Ga., at commencement rcises on June 5, 1949. The degree is being ferred for distinguished service as an icator, engineer, and practical scientist l is the highest honor the school can be- w. Mr. Head is president and founder of Electronics Institute, Inc., of Detroit, , and also directs the activities of In- trial Electronics, Inc., a professional sultant organization.

A graduate of Oglethorpe University in 5, Mr. Head later became associated with e DeForest, the "father of radio." He s recently elected president of the De- it Section of the Instrument Society of erica, and holds membership in the De- it Television Roundtable and the Engi- ring Society of Detroit.

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Roger E. Robertson (A'48) has been ap- nted to the staff of the National Bureau Standards, where he will conduct engineer- research for the guided missile projects he electronics division.

A native of Jaffrey, N. H., Mr. Robert- won the freshman competitive scholar- to the Massachusetts Institute of hnology, and was the recipient of further dergraduate and graduate scholarships. ile studying for the bachelor of science master of science degrees in electri- engineering, both of which he received 1942, he worked on co-operative assign- nts at the General Electric Co. Upon his duation he joined General Electric on a -time basis, leaving in 1946 to become an tronics engineer at the Bell Telephone p., where he directed and supervised the tronic phases of a flight-test program for ded missiles.

Mr. Robertson has worked on the design ire-control radar and other electrical and tronic systems. He is a member of Eta ppa Nu.

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Elmer W. Engstrom (A'25–M'38–F'40) e-president of RCA in charge of research s given the honorary doctor of science de- e by New York University on June 15, 9.

George C. Schleter (A'38–VA'39) has been appointed to the staff of the National Bureau of Standards, where he will conduct an engineering development program on guided missiles, including missile systems and components, in the Bureau's electronics laboratories.

Mr. Schleter was born in Seymour, Ind., and attended Purdue University, from which he received the B.S. degree in 1922. He spent the next three years as a member of the faculty of New York University, from which he received the master of science degree in 1925. In that year he was appointed head of the mathematics and physics departments of Broaddus College in Philippi, W. Va., and he left in 1928 to become a civilian physicist on the staff of the U. S. Army Air Corps at Wright Field.

From 1931 to 1934 Mr. Schleter worked on the development of blind landing systems at the National Bureau of Standards; then he engaged in microwave research at the Naval Research Laboratory, as well as developing antennas for ASB radar, radio transmitters, and confidential microwave research. Early this year he rejoined the National Bureau of Standards.

Mr. Schleter is a member of Eta Kappa Nu.

❖

Everard M. Williams (S'36–A'41–SM'44), of the Carnegie Institute of Technology, has just been promoted to a full professorship at that institution.

Born in 1916, Dr. Williams studied at Yale University, receiving the doctorate in 1939. Shortly afterward he became an instructor at Pennsylvania State College, leaving in 1942 to become chief engineer of a secret radio weapons laboratory at Wright Field, Dayton, Ohio.

At the end of the war, in 1945, Dr. Williams joined the faculty of the Carnegie Institute. There he helped in the development of the new Synchro-Cyclotron, of which he designed the oscillator and deflector. In 1946 he was voted America's outstanding young electrical engineer for 1946 by the Eta Kappa Nu Assn. He was also awarded the President's Certificate of Merit for his war work.

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Paul Ware (A'17–SM'44) has been appointed head of the DuMont Laboratories' new electronics parts division.

From 1907 to 1912 Mr. Ware served as a wireless operator for Atlantic Deforest, American Deforest, United Wireless, and Marconi. He received the M.E. degree from the Stevens Institute of Technology in 1917 and subsequently enlisted as a sergeant in the U. S. Army during World War I. Rising to the rank of second lieutenant, he was responsible for the 75-meter break-in trench telegraph set known as SCR77, on which he had previously worked with Professor Hazeltine at Stevens.

From 1922 to 1925 Mr. Ware was engaged in operating one of the pioneer radio receiver enterprises which developed, manufactured, and sold over 200,000 receivers known as the "Ware Neutrodyne." He developed and patented a duplex radio communications technique embodying the break-in feature and many other improvements important at the time, including the first completely-shielded receiver and the double-oppositely-wound loop.

From 1925 to 1935 Mr. Ware was consultant for Splitdorf-Bethlehem, Sonora Phonograph, R. E. Thompson, and also for the Monmouth Memorial Hospital, and engaged in the design and manufacture of radio sets for various companies. In 1935 he joined P. R. Mallory and Co. as consulting engineer and remained there until 1939, when he transferred to DuMont.

In 1942 and 1943 Mr. Ware served as president of the Radio Club of America.

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Leslie J. Woods (M'35–SM'43) has been appointed vice-president and director of research and engineering to head all engineering and research activities of the Philco Corp.

Mr. Woods joined Philco in 1925, after serving in the British Army during the first World War, and playing an important role in the development of British communications in the Middle East. In 1928, when Philco began research work in television, Mr. Woods was named first television engineer, and served in this capacity for two years. From 1930 to 1938 he occupied positions of increasing responsibility in the company's engineering department, designing and developing both vacuum tubes and radio receiving sets. During this period he took charge of automobile radio engineering for the corporation, and in 1941 was made manager of the car manufacturers' division, with headquarters in Detroit.

On the outbreak of war he was transferred to Washington, where, in 1942 he became vice-president and general manager of the National Radio Corp., at that time a Philco subsidiary, to assist that company in expanding its organization to meet the greatly increased wartime demands of the army and navy. Following the war, he returned to Philco as manager of the company's industrial division, which takes care of automobile and aircraft radio, as well as advanced radar equipment for the armed forces, and in 1948 he was elected vice-president of the division.

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Daniel R. Donovan (A'44), formerly vice-president and sales manager of the Callite Tungsten Corp., has been appointed sales manager on elmet and fine wire products, for the North American Philips Co., Inc., in Lewiston, Me.

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George L. Downs (A'44) has been placed in charge of the transformer operation of Ratheon Manufacturing Co., at Waltham, Mass. Mr. Downs was formerly head of Raytheon's production test department, and methods and cost estimating department. A native of Texas, he is an amateur radio operator, his call letters being WICT.

Cover Sheet for Technical Memoranda—A Technique in Information Exchange*

R. C. MATHES†, SENIOR MEMBER, IRE

Summary—**An account is given of the origin and use of a routine for disseminating new technical information and ideas rapidly and flexibly through the large engineering and research organization of Bell Telephone Laboratories. The essential philosophy back of this routine is that technical memoranda have the status only of the individual engineer's personal authority, and hence may be circulated directly and widely across departmental lines for information and discussion. Departmental policy in relation to such memoranda is established by covering letters where such action is necessary. The objectives approached and benefits accruing from this procedure are briefly discussed.**

Introduction

ONE OF THE routines to which a new technical employee in Bell Telephone Laboratories is soon introduced is that of the "Memorandum for File," together with its inseparable companion, "Cover Sheet for Technical Memoranda." Such an individual may be tempted to regard this latter as a piece of bureaucratic red tape designed to supply work for stenographers and filing clerks, but in time he learns that it has definite relations to the organic needs of our type of work.

Starting in a small way about twenty years ago in a small research group working on carrier transmission, this cover sheet has had its use spread in turn to the department, the general department, and finally to all of our technical general departments. We will not trace out the little variations in its form and use from then to now, as a description of its present status will suffice to bring out the essential philosophy in back of its initiation and survival.

Without attempting to delve deeply into the nature of our Laboratories' organization or its technical work, there are two aspects of it which appear apposite to our present discussion. First, there is a vertical division of effort between such major functions as Research, Apparatus, and Systems. Each of these, in turn, is divided into other functional divisions, such as Physical Research, Chemical Research, Acoustic Research, Transmission Research, etc., which in turn are broken down into still more specialized compartments. It is in these specialized compartments that we strive for new knowledge, new types of apparatus, and new systems for application. The second aspect to our method of working is a horizontal one. Closely related lines of work must continually draw on each other for supporting information and specific projects, such as a new machine switching system or a microwave transmission system will need to be in touch with the work of a wide range of specialized workers. Within this simplified picture of the duality in methods of operation, it can be seen that there are numerous areas of work which need to be co-ordinated, continuously and flexibly. The technique to be described is one means to this end.

Objectives

From the above brief introduction to the nature of the problem, a little consideration will enable us to set up a group of objectives whose attainment we believe to be furthered by the use of the cover sheet.

1. Rapid dissemination of information to those most probably interested.
2. The encouragement of direct contact between individual engineers.
3. The avoidance of extensive duplication of effort.
4. The avoidance of vested interests in or undue isolation of fields of work.
5. Speeding up reviews of past work.

The two media which we have for the spreading of ideas are the oral and the written word. From the standpoint of speed and fluidity of interchange of ideas the former is the more effective, while the latter has its advantages from the standpoints of record and study. In the Laboratories, the use of the oral word is fostered by the medium of conferences in groups and between individuals with all the advantages attendant upon personal acquaintanceship. However, in a large organization one can never be quite sure that all necessary or desirable personal contacts are established, and time alone precludes trying to establish all possible ones. This situation is rendered more serious by the fact that many, if not most, of the best engineers and research men are of the subjective type,[1] and do not venture far in establishing personal contacts.

The Mechanism

It thus becomes important to have a method for using the written word in the most flexible and direct fashion for furthering the objectives outlined above. The cover sheet, a typical example of which is shown in Fig. 1, has been evolved to supply this need. The first line is the subject and the case number under which expenditures for the work are authorized. Below this on the left is a routing list. If the memorandum is deemed to contain patentable material, the No. 1 copy, signed by the author, is forwarded to the Patent Department, generally with a covering letter containing additional discussion. The second, or yellow copy, also signed, goes to the Case File for permanent record. The third copy generally goes up the line of supervision to the department files. It is th mainder of the routing list which acc plishes our object of rapid information semination. Here the author lists the na of individual engineers in his own or o departments to whom he believes the sub is of direct interest. There may be only a or there may be several dozen. If he beli that a certain group is or ought to be ir ested, but he does not know the individ concerned or their responsibilities, he route a copy to the group head. The engi may confer with his immediate supervis to the general coverage of the list. It cut across not only all department lines, also all ranks from a general department h to individual engineer.

To the right of the page is a routine identification number, the date and the thor's name. MM signifies "Memoranc for File." The identification number con simply of the last two digits of the year, department number, and the serial nun of the memorandum for that departmen that year. Under the author's name the included a filing subject or type of work t

The objective of rapid disseminatio ideas and knowledge is attained by virtu the way in which the routing list cuts ac all departmental lines. No time is lost in ing five or six steps up and a similar num down with the inevitable delays due to p sure of other work, questions raised, etc. objective of encouraging personal conta attained, because an individual of the s jective type (who may rarely speak up group conference) will, upon reading a port of special interest to him, reach for telephone to make an appointment for a sonal conference. His object may be to further information or to argue a diver viewpoint, but either way a new friend or improvement in mutual interests ge ally results.

Below the routing and filing informa on the cover sheet there is a brief abst stating the substance of the memorand This is of especial interest to supervisors department heads. First, in relation to jective No. 1 they can review the routing to see if every one in their group who sh have a copy of that material is on the Second, it enables them to keep a run check on the third and fourth objectives ing to do with interdepartmental divisio effort and interdepartmental correlation

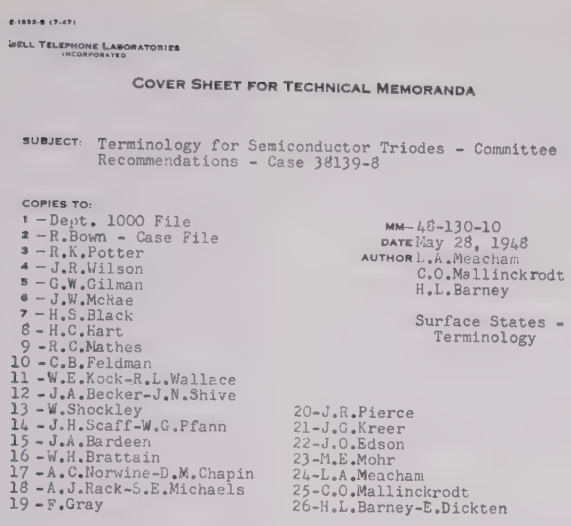
The fifth objective is furthered if a fi kept by subject matter. We have done in the past by keeping a file of cover sh only arranged by subject or class of work present this type of file is being kept on inch cards. An engineer reviewing or reo ing a line of work thus has available a brief guide to the more important me randa to read. He also gets a good steer which individuals he should seek out firs personal contact.

A further check on the adequacy of

* Decimal classification: R009. Original manuscript received by the Institute, March 3, 1949.
This paper is an expansion of discussion prepared in relation to the paper by Allen H. Schooley, U. S. Naval Research Laboratory, entitled, "Information exchange as a management tool in a large research organization," presented at the National Electronics Conference, Chicago, Ill., on November 5, 1948, and published in PROC. I.R.E., vol. 37, pp. 429–432; April, 1949.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] C. E. Broadly and M. Broadley, "Know Your Real Abilities," Chaps. 6 and 7, McGraw-Hill Book Co., New York, N. Y. 1948.



BELL TELEPHONE LABORATORIES
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COVER SHEET FOR TECHNICAL MEMORANDA

SUBJECT: Terminology for Semiconductor Triodes - Committee Recommendations - Case 38139-8

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MM - 48-130-10
DATE May 28, 1948
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H.L.Barney

Surface States - Terminology

ABSTRACT

Recommendations are made for an equivalent circuit representation, and terminology relating to semiconductor triodes.

Fig. 1

ution is maintained by the monthly circu-
n, to a certain level of supervision, of a
of titles and authors of all technical
noranda for file. Several sets of circula-
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ence basis to take care of requests result-
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Status of Technical Memoranda

It may be objected that this free circulation of memoranda might induce confusion between the work of departments. This is avoided by the status accorded these memoranda. They are regarded as the technical expositions of his work by the individual engineer and carry only his personal authority. Thus, a supervisor may approve[2] them for file and circulation, even when not in full accord with some of the ideas expressed. They are only made a part of interdepartmental policy and action when transmitted with covering letters signed by supervision and to the degree set forth in each letter.

From this independent status given to the technical memoranda for file, there derive benefits quite aside from the original objectives of dissemination of information. These memoranda become, for the individual engineer, a medium of publication to a large group of his associates, not only of the results of work done, but also of his judgment of the significance of the work and of his plans and hopes for the future. Sympathetic supervision can thus encourage the development of the individual through constructive criticism of this medium of expression. The high standard attained in the writing of many of these memoranda is testified to by the fact that, not infrequently, they become, with only slight changes to adapt them to a new and wider audience, the manuscripts for publications in the technical press.

[2] There is some feeling that even this degree of approval is not necessary.

Radioactive Standards and Methods of Testing Instruments Used in the Measurement of Radioactivity*

L. F. CURTISS†

ummary—**A summary of the program for standation of radioisotopes at the National Bureau of ards is given in this paper, with some of the ns for need of absolute measurements of radioes in particular uses. This is accompanied by an e of some of the methods used to ascertain charstics of commercial instruments used in the rement of radioactivity, with particular empha those methods used at the National Bureau of ards for this purpose.**

Introduction

THE MEASUREMENT of radioisotopes has assumed new importance in recent years. This can be attributed he increased production of artificially produced isotopes, which, in turn, has led to a greater diversity of uses. The expansion in the requirements for measurements has developed much faster than the development of equipment for the purpose, both in quantity and in quality. Instruments commonly used in measurements of radioactive radiations rarely have been of the type which could be reliably calibrated quantitatively with the expectation that such calibrations could be relied upon over long periods of time. As more sensitive and complex instruments have been developed, the instability has increased. Therefore, many difficulties are encountered, which are chiefly of an instrumental nature, where absolute measurements of activity are required. While this situation may be said to result indirectly from the increased amounts of various radioisotopes now available, it arises chiefly from new uses which have been found for these isotopes, where absolute measurements are essential.

Before radioisotopes were available in quantities to encourage practical uses, they were mainly the subject of investigations regarding their structure, modes of disintegration, and related basic information. Measurement in this field can be accomplished without any knowledge of the absolute amount of the isotope under study. Each investigator sets up his own arbitrary scale of units, based on instruments and techniques usually available only to himself. The only comparison of his results with those of others, using different quantities and methods, is made in the conclusions deduced from the measurements. In many tracer experiments

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National Bureau of Standards, Washington,

a similar situation exists. An arbitrary scale of activities set up for a particular experiment suffices to yield the desired quantitative results.

There is one radioisotope which has required quantitative measurements for many years. This radioisotope is radium. Accurate standards[1] of radium were set up internationally in 1913. A survey of the principal reasons for this procedure throws considerable light on the nature of the present demand for absolute measurements for artificially produced radioisotopes. The more important of these are that (1) radium has become an article of commerce; (2) it is used principally in the treatment of disease; (3) certain physical investigations required absolute measurements of this isotope. An example of (3) is the determination of the disintegration rate of radium.

The same situation regarding radium is now true of many of the artificially produced radioisotopes, and the need for their absolute measurement has the same general basis. This has created a demand for standards of a large number of radioisotopes in forms suitable for calibration of a wide variety of instruments. The more recently developed instruments for radioactive measurements are more sensitive than the earlier instruments. They also exhibit a greater degree of instability, and have more complex components which may fail to function properly. Therefore, the usual procedure in measurements of radioactivity—that of frequently checking the response of an instrument by using a standard—has become more necessary with present-day instruments. Most of these instruments are linear in their response only over limited ranges. This necessitates a series of standards of different intensities to reduce errors which might be caused by this lack of linearity in the instruments.

Accurate standards will not, by themselves, solve the problem of accurate absolute measurement. It is obviously equally important that the instruments function reliably within their limitations, and that these limitations be clearly recognized. As the instruments have increased in complexity, as compared with the gold-leaf, or Lauritsen, electroscope, with which many important measurements have been made, the number of tests which must be made to ascertain proper operation has increased. The testing of instruments is thus a new requirement in radioactive measurements. Until quite recently, practically all equipment used in radioactive measurements was constructed in the laboratory where it was to be used, usually by the investigator himself. Thus, the construction and operational characteristics were entirely familiar to the user. This situation no longer obtains. Measuring equipment is now being produced commercially. Numerous companies are engaged in this production, with no two companies making a given instrument in the same way. The prospective user must, therefore, before he uses a particular instrument, ascertain whether it operates properly and determine its peculiar limitations. This must be accomplished either by tests which he makes himself or which are made for him. There is a sufficient diversity in instruments made by a single manufacturer to necessitate individual tests for each instrument. In many situations these tests must be made for the user, who frequently may not have the equipment and training to make them for himself. Thus we see that standards of radioactivity, and tests of instruments with which these standards are used, are both essential to absolute measurements of radioactivity.

It is the purpose of this paper to describe some of the standards of radioisotopes already available; to outline the program for producing additional items in this list, and to discuss some of the precautions required in the correct use of these standards. This is followed by a description of some of the tests required for a few types of instruments, including the equipment used for this purpose.

Standards of Radioactivity

A radioactive standard is conveniently defined as a specimen of a given radioisotope for which the amount of the isotope present has been determined quantitatively. An illustration is provided by the National Radium Standard which is deposited at the National Bureau of Standards. In 1913 Madame Pierre Curie prepared some very pure radium salts which were used to determine the atomic weight of radium. Several hundred milligrams of these salts were used. After the work of determining the atomic weight was completed, this material was divided into a number of portions, each portion carefully weighed and sealed in a glass tube. These were then primary standards of radium, since the number of milligrams of the isotope radium in each tube had been accurately determined. This method of preparing a standard could be used for radium since the amount of radium, in equilibrium with its decay products, which produces a conveniently useful amount of radiation is of the order of several milligrams. This is a direct result of the slow rate of disintegration of radium.

A little consideration shows at once that the gravimetric method of determining the amount of radioisotopes will fail for those isotopes which disintegrate more than 100 times as fast as radium. We might consider Co^{60} as an example. A sample of radium containing 10 mg of the isotope will have 2.66×10^{19} atoms of the isotope. The number of atoms disintegrating per second will be this number multiplied by λ, the decay constant for radium. This constant has the value 1.36×10^{-11} per second. Hence, the disintegration rate per second for 10 milligrams of radium is $2.66 \times 10^{19} \times 1.36 \times 10^{-11} = 3.61 \times 10^{8}$. We will now compute the weight of an amount of Co^{60} which has the same rate of disintegration. The decay constant for Co^{60}, assuming a half-period of 5.3 years, has the value of 4.15×10^{-9} per second. The number of atoms required is, therefore,

$$\frac{3.61 \times 10^{8}}{4.15 \times 10^{-9}} = 8.70 \times 10^{16} \text{ atoms.}$$

Since there are 10^{22} atoms in a gram of cobalt, the weight of this amount is

$$\frac{8.7 \times 10^{16}}{10^{22}} = 8.7 \times 10^{-6} \text{ grams.}$$

A microchemical balance of highest sensitivity, used under the most exacting conditions would be required to determine this weigh to an accuracy of a few per cent.

Co^{60} has a half-period of 5.3 years. Many radioisotopes in use today have half-periods of a few days. The gravimetric method is ou of the question for them. This leads to a consideration of radioactive methods of determining the amount of a radioisotope. We already have an example of this in the members of the decay chain of radium. For example, polonium has a half-period of approximately 140 days. Although it cannot be weighed, its rate of disintegration can be determined by measuring the rate at which alpha particles are emitted. The amount of polonium in equilibrium with 1 gram of radium, in analogy to the case of radon, is defined as 1 curie of this isotope. The rate of disintegration of 1 curie is the same as for 1 gram of radium; that is, 3.61×10^{10} disintegrations per second. Hence, the number of millicuries in a preparation of polonium can be ascertained by measurements of the radioactivity in terms of rate of disintegration If necessary, the number of atoms present can be computed by dividing this rate by the disintegration constant.

In principle, this method of measurement may be extended to any radioisotope. However, the details become more complicated when applied to those isotopes which do not disintegrate by the emissoin of alpha particles. Those isotopes which emit beta rays or positrons, or which disintegrate by electron capture, frequently have complicated disintegration schemes which do not permit a simple computation of the disintegration rate from measurements which can be made of the radiations emitted.

Even in the simple case of an isotope which disintegrates only by emitting beta particles of a single maximum energy, the nature of the continuous spectrum of nuclear beta rays introduces a difficulty. These beta rays are emitted with all energies from zero to the maximum. Any convenient detector for measuring the rate at which they are emitted will fail to detect those below a certain minimum energy. This energy depends on the characteristics of the detector and the amount of absorber between the source and the active volume in which the particles are detected. Furthermore, these particles are emitted with an equal probability in all directions. This requires an accurate determination of the solid angle subtended by the detector window at the source. Proceeding to the more complicated modes of disintegration, where two or more beta rays of different maximum energies are emitted, each followed by one or more gamma-ray transitions before the stable, nonradioactive state is reached, other difficulties appear.

One of the most serious is that some of the gamma rays may be internally converted. This process consists in the ejection of a shell-electron from the atom by a gamma ray from the nucleus. Therefore, in place of a gamma ray, an electron is emitted with an energy equal to that of the gamma ray decreased by the binding energy of the electron in its shell. These conversion electrons thus have energies which are usually comparable with those of individual beta rays. In ordinary measurements with counters, they can

[1] See definition in section on Standards.

be distinguished from beta rays. This s to erroneous determinations of disinte- ion rates, unless an accurate correction ade for the conversion electrons.

t is obvious that disintegration rates ıot be determined unless all details of the ıtegration scheme are available. Fortu- ly, the number of radioisotopes for ch disintegration schemes are known is dily increasing. At present, however, represent a small fraction of the known pes which might have practical use.

The general situation is such that the av- e worker who needs to measure radioiso- s in absolute terms might well be ouraged. His difficulties would be ved if he had standards for each isotope e measured. The measurements would consist in mounting the sample to be sured in any convenient way with re- t to the detector, and observing the t. A standard of suitable strength nted in the same way would give the to compute the disintegration rate of sample. This procedure transfers many ıe complications in regard to the meas- ıent of disintegration rates to the labo- y which prepares the standards.

The short half-lives of many of the arti- ly produced radioisotopes do not per- the preparation of permanent or semi- ıanent standards. These can be pre- d for such isotopes as Co^{60} and Na^{22}, e the half-life is of the order of years. those isotopes having half-periods meas- in days, resort must be made to stand- ing expendable samples which are ibuted by a standardizing laboratory at ciently frequent intervals to maintain alibration of the measuring equipment.

Calibration of Standards

Various methods are available for the ration of standards. We will discuss typ- xamples of the more important of these ustrate the principles.

ne of the most reliable methods of de- ining absolute disintegration rates is the idence method. Perhaps an explanation e meaning of the term "coincidence" in connection is justified. By referring to , which shows the disintegration scheme ^{60}, we note that the disintegration oc- by the emission of a single beta ray n transforms the nucleus into Ni^{60}, and the nucleus of the newly formed nickel emits two gamma rays before it be- s stable. The nucleus is then said to be in ound state, which means that it has no nal energy in excess of that possessed by able nonradioactive nucleus. Actually, must be a finite interval of time be- n the emission of the beta ray and the quent emission of the gamma rays, we know that the shell-electrons in the formed nickel atom have had time to ge themselves in the levels appropriate his atom, which differ distinctly from vels of the cobalt atom. However, this val of time is too short to measure by conventional methods, and is probably e order of 10^{-12} second. Consequently, l practical purposes the gamma rays in- aneously follow the beta ray with which are associated. This means that, under able conditions, it should be possible to

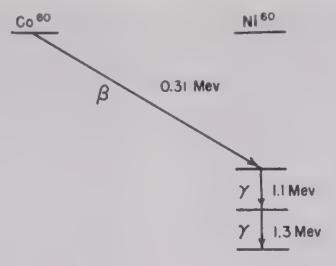


Fig. 1—Disintegration scheme for Co^{60}. In this disintegration a single beta-ray spectrum is emitted with a maximum energy of 0.3 Mev. Each newly formed nucleus of Ni^{60} emits two gamma rays, immediately bringing the nucleus to the ground state.

detect a gamma ray which is emitted simultaneously, within the limits of the resolving power of usual arrangements for detection, with a beta ray. This supposition has been confirmed. The determination of the relative number of these simultaneous events is called a coincidence measurement. In addition to beta-gamma coincidences, it is obvious that gamma-gamma coincidences occur in the disintegration of Co^{60} and, in principle at least, it should be possible to observe them. The disintegration of Co^{60} follows what is called a simple disintegration scheme, since it involves only one beta ray. In more complex schemes the details can frequently be unravelled by a study of the coincidence rates between the various radiations emitted. The coincidence method can only be used for those isotopes emitting beta rays followed by one or more gamma rays. It is most reliable when no measurable amount of the gamma rays suffers internal conversion.

Consider an arrangement in which a beta-ray counter and a gamma-ray counter, respectively, detect beta particles and photons from the same source. These counters are connected to electronic equipment which will record the photons and beta particles independently. It also records the coincidences between the two counters. Fig. 2 shows a block diagram of the equipment. The observations are corrected for counter backgrounds, chance coincidences, and cosmic-ray coincidences. Let N be the actual disintegration rate of the source, A the rate of the gamma counter, and B and rate of the

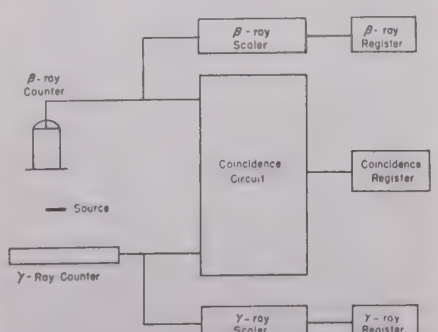


Fig. 2—Block diagram of a coincidence circuit for determining β-γ coincidences.

beta counter; E_γ represents the over-all efficiency of the gamma-ray counter, and E_β that of the beta-ray counter. For present purposes we need not know the actual value of these efficiencies, which are defined as the fraction of the total number of particles or photons emitted which are detected by the respective counters.

Then we have

$$A = NE_\gamma \quad \text{or} \quad E_\gamma = \frac{A}{N}$$

$$B = NE_\beta \qquad E_\beta = \frac{B}{N}$$

If C is the coincidence rate,

$$C = NE_\gamma E_\beta = N\frac{A}{N}\frac{B}{N},$$

whence

$$N = \frac{C}{\frac{A}{N}\frac{B}{N}} = \frac{N^2C}{AB}$$

or

$$N = \frac{AB}{C}.$$

Thus the absolute disintegration rate is obtained by dividing the product of the rates of the rates of the two counters by the coincidence rate. The efficiencies do not appear in the final result.

For an isotope which emits only beta rays, the absolute disintegration rate can be obtained by counting all particles in a definite geometry under conditions where all absorption and scattering effects have been reduced to a negligible amount. The simplest geometry is that which includes all particles in the 4π solid angle about the source. In practice, this is a little difficult to obtain, since the source must be supported. The effect of the support can be reduced by making it very thin. One satisfactory way to make measurements of this type is to use a large evacuated chamber with the source mounted on a very thin support at the center. The rate at which the source and support acquire electric charge is then a measure of the disintegration rate. This requires moderately strong sources for accurate results. A modification of this principle of using a definite geometry is to set up a series of apertures which define a small solid angle, which can be computed from the dimensions. The region in which the beta rays travel to the detector is evacuated. If the detector has a window of a thickness less than 0.5 mg/cm^2 and the maximum energy of the particles is of the order of 1 Mev, only a small fraction of the particles emitted in the defined angle will miss detection as the result of absorption. This method is not very satisfactory for beta rays whose maximum energy is much less than 1 Mev.

The problem of calibrating standards in terms of disintegration rates becomes more difficult when we deal with radioisotopes which emit beta rays in two or more continuous spectra with different maximum energies, each followed by several gamma rays. A very careful study of the disintegration scheme is required in order to devise a coincidence arrangement which will give meas-

urements which can be interpreted in terms of disintegration rates. This problem is further complicated when some of the gamma rays undergo internal conversion. The conversion electrons usually will be of sufficient energy to be detected in the beta-ray counter. The relative number of these conversion electrons must be determined, and a correction made for their contribution to the particles recorded by the beta-ray counter used.

The problem of producing standards for radiosotopes has become much more complicated than that of preparing standards of radium, as far as physical measurements are concerned. The chemical balance has been replaced by instruments for determining the rate of disintegration of the sample undergoing standardization by measurements on the radiation emitted. Since many radioisotopes do not have half-periods which permit the preparation of permanent standards, the production of standards is a continuous process which must be repeated periodically.

The demand for radioactive standards is naturally greatest for those radioisotopes which are used in relatively large amounts, under conditions where absolute measurements are required. A tentative list of such radioisotopes is given below.

	Half-Period
C^{14}	about 6000 years
Na^{24}	14.8 hours
P^{32}	14.3 days
S^{35}	87.1 days
Ca^{45}	180 days
Fe^{59}	42.5 days
Co^{60}	5.3 years
I^{131}	8.0 days

All of the isotopes in the above list can be produced in a pile. Thus they can be made available in large amounts, which accounts in part for their extensive use. If radioisotopes are produced in comparable quantities in cyclotrons, there are a number of other radioisotopes which may become widely used. The most important of these, at present, is Na^{22} (3.0 years).

C^{14} and I^{131} are used extensively for biological and medical investigations, and standards of these isotopes are therefore needed. The National Bureau of Standards has a program for the development of standards for C^{14}. At present it appears that these standards will be most useful if prepared as calibrated expendable samples, rather than as permanent standards. Although the long half-life would permit permanent standards, it is difficult to prepare samples of compounds containing radioactive carbon in which the ratio of the active to the inactive isotope remains constant. For example, solid deposits of barium carbonate might be used as standards. In use these would be exposed to the atmosphere, and it is known that such samples exchange CO_2 with the atmosphere in the presence of moisture. Therefore, they would gradually lose C^{14} atoms with use, even if stored in a dry atmosphere when not in use. The problem of providing specific forms of standards for all types of possible uses becomes so complicated that some simplification is required. Some investigators need to use the standard in the form of CO_2, and others need to prepare their working standards to conform to the requirements of a specific type of measuring equipment. For the present, it is proposed to furnish these standards in the form of calibrated solution of sodium carbonate sealed in glass ampoules. The calibration will give the disintegration rate per milliliter, so that definite amounts of this solution can be removed and converted into a working standard of the desired form.

In the case of I^{131}, the short half-period precludes the preparation of permanent standards, and calibrated expendable samples must be used. Although I^{131} emits both beta rays and gamma rays, and therefore should be susceptible of calibration by coincidence methods, the disintegration scheme is complicated by the presence of two beta rays and a number of gamma rays, some of which undergo internal conversion to a measurable degree. When the details of this scheme are completed, a satisfactory coincidence method for calibration can be made available. Present information allows calibration to ± 5 per cent.

The National Bureau of Standards has made several distributions of identical samples of I^{131} over the past two years. The reported results reveal a disagreement in measurements approaching a factor of 2. These disagreements are to be attributed, in part, to the lack of suitable standards with which to compare the activity of samples of I^{131}. It has been indicated that the use of a standard of one isotope to calibrate a quantity of a different isotope may lead to serious errors. Therefore, if one attempts to measure I^{131} with a RaD+E standard which has a very different beta-ray spectrum from that of iodine, a considerable number of corrections and precautions are required to obtain results that are not in error by more than 20 per cent.

It is, therefore, obvious that many of the possible sources of error can be eliminated, if calibrated samples of I^{131} are used for the measurements. Using such samples, the results should be free from serious errors if the unknown and calibrated samples are compared under identical conditions. If this requirement is complied with, no corrections need to be applied to the data used in computing the disintegration rate of the unknown sample. This reveals the need for a distribution of calibrated samples of this isotope at intervals of the order of every few months. Work is in progress to prepare and distribute such samples.

P^{32} emits only beta rays in a single spectrum with a maximum energy of approximately 1.7 Mev. This isotope therefore cannot be calibrated by coincidence measurements. The most convenient method of calibration is counting the beta particles in definite solid angle with corrections for absorption and scattering. Those using P^{32} can make a fairly accurate measurement of quantities deposited by evaporation, using the RaD+E standards prepared by the National Bureau of Standards. These standards contain a measured amount of RaD. Within a month of perparation, RaE grows into equilibrium with the RaD, so that the disintegration rate of the RaE is known. Fig. 3 shows one of the standards and two blank disks on which standards are deposited. The RaD is deposited on the palladium face of a silver disk 1/16 inch thick and 1 inch in diameter, faced with a layer of palladium

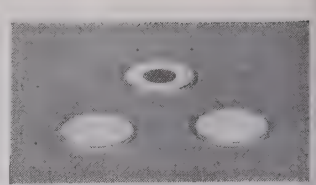


Fig. 3—RaD(Pb^{210})+RaE(Bi^{210}) standa
Two of the disks are blank and the t
has a circular deposit of lead, par
which is Pb^{210}. Approximately one m
after preparation, RaE is in equilib
with the RaD, after which the acti
decays with the half-period of
(22.2 years). Blank silver disks are
to mount deposits for comparison o
tivity with the standards.

0.002 inch thick. The deposit is circular
approximately 12 mm in diameter. It i
cated centrally on the face of the disk.
disk provides saturation back-scatterin
the beta particles. If the phosphorus is
posited on a similar disk the back-scatte
correction is reduced to a few per cent. T
RaD+E standards can be used for
erately accurate measurements of isot
which disintegrate only by beta ray or
tron emission when the spectrum is si
and has a maximum energy near that o
RaE spectrum. Appropriate absorption
rections must be applied in each case.
absorption correction is determined ex
mentally at the time of comparison by
trapolating the beta-ray absorption cur
zero thickness of absorber.

It should be noted that these stand
emit soft beta rays from RaD, and a
particles from the polonium which is g
ing in them. Therefore, they must alway
used with a filter which excludes these r
tions from the counter. A layer of alumi
0.001 inch thick is adequate for this
pose.

Fig. 4 shows a series of radium gam
ray standards consisting of 5 millilite
solution sealed in Pyrex ampoules. Ran
from 0.1 micrograms to 100 microgr
they are intended for gamma-ray mea
ments of radium. Since the solution doe
completely fill the ampoule, radon col
in the air space above the solution, pro
ing an uneven distribution of activity.
disadvantage can be overcome by sha
the solution vigorously several minutes
to use.

Radium in amounts less than 10^{-8} g
is most accurately determined by meas
the production of radon from the radium
in solution. The standards required for
measurement consist of a solution conta
a known amount of radium. Fig. 5 show
standards and a blank solution. The ra
is contained in 100 milliliters of solution
standard having a total radium conte
10^{-9} grams, the other 10^{-11} grams. Near
water, no matter how carefully distilled,
tains measurable amounts of radium. T
fore, ordinary distilled water cannot be
to rinse the flasks containing the stand
when transferring the solutions to th
paratus in which it is to be used. To
might involve a risk of adding severa
cent to the amount of radium in the s
ard. This difficulty is removed by the

. 4—Gamma-ray standards. Microgram standards of radium. The standards cover a range from 0.1 to 100 micrograms of radium, and are used for gamma-ray measurements of radium.

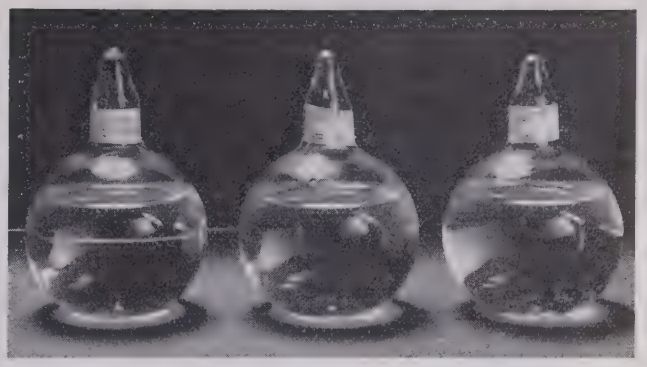
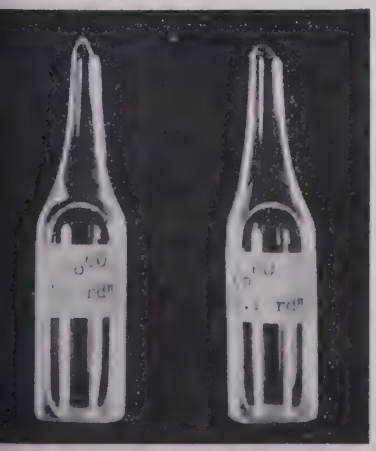


Fig. 5—Radon standards. These flasks contain 10^{-11} and 10^{-9} grams of radium in 100 milliliters of solution. The blank solution has a measured low radium content, and is used for rinsing when transferring the standard solution to the apparatus in which it is to be used.

ank" solutions sealed in flasks similar to standard. The amount of radium in these utions has been measured and found to be 5×10^{-12} grams of radium. This amount is gnificant compared with 10^{-9} grams of ium in the larger standard, but it must be en into account when used with the 10^{-11} m standard.

The only other isotope for which the Naal Bureau of Standards has prepared brated standards at present is Co^{60}. This tope lends itself readily to the coincidence thod of calibration. The current standrs are gamma-ray standards, since they sist of a 5-milliliter solution sealed in a ss ampoule. The solution from which se standards were prepared was measured hree different laboratories by the coincice technique with agreement in the rets of the order of 1 per cent. They are ilable in two sizes; 1.5 rutherfords and 5 rutherford. Typical ampoules are shown Fig. 6.

The rutherford (rd) is a unit recomnded by the National Bureau of Standrs for expressing quantities of radioisoes other than members of the radium nily. It is defined as the amount of a raisotope undergoing 10^6 disintegrations per ond. It is, therefore, a decimal unit which facilitates computations. It is also of a convenient magnitude for expressing quantities of a radioisotope usually handled and measured in the laboratory. For example, a sample which can be conveniently measured by an end-window Geiger-Müller counter is of the order of 0.1 to 0.4 millirutherford. Tracer quantities are of the order of 1 rutherford. Therapeutic doses are of the order of 100 rutherfords. In addition to covenience there is another reason for the introduction of a unit which is uniquely defined in terms of disintegration rates. This is the widespread misuse of the curie. It has become fairly common to use the curie to refer to quantities of radioisotopes when the disintegration rate has not been measured. It is even used frequently when disintegration rates cannot be measured. It has been pointed out that a complete knowledge of the disintegration scheme is necessary before a disintegration rate can be measured by radioactive methods. Therefore, when we see quantities of isotopes expressed in curies for which disintegration schemes are either partially or wholly unknown, we can only conjecture as to the quantities intended. Generally, such statements have no meaning and only lead to confusion. Finally, there is still considerable uncertainty regarding the value of the curie when used as a unit disintegration rate. Logically, from its original definition as "the quantity of radon in equilibrium with 1 gram of radium" it has the value of the disintegration rate of 1 gram of radium. Values for this figure varying from 3.4 to 3.8×10^{10} have been reported by competent investigators. The most recent value is 3.61×10^{10}. The value 3.4×10^{10} is still in current use. An attempt to smooth over this situation has been made by adopting 3.700×10^{10} as an arbitrary figure. However, recently Madame Joliot-Curie has proposed an "international curie" to have the arbitrary value 3.600×10^{10}. The result of this situation is that it is incumbent on each author of a paper to state the value of the curie which he uses. The rutherford eliminates this difficulty, since it is not associated with any natural constant, the value of which may fluctuate with the precision of measurement, or the arbitrary selection of different values by different authors.



. 6—Co^{60} standards. The cobalt solution is sealed in a glass ampoule which then serves as a gamma ray standard for measurement of solutions of Co^{60}.

Testing of Instruments for Radioactive Measurements

Instruments used in the measurement of radioactivity may vary from the simple gold leaf or quartz-fiber electroscope to the complicated assembly of electronic equipment required to operate a Geiger-Müller counter and record the pulses from it. Devices which use ionization chambers as the sensitive element also frequently employ electronic circuits to detect the required direct-current potentials produced in the measurement of the ionization currents. In the more complex assemblies, over-all checks may reveal either satisfactory or unsatisfactory operation, but in cases where performance falls short of that expected or required, detailed tests must be made of each component. In any case, a laboratory testing equipment of this type must be prepared to test the various components, since they are frequently purchased separately from different manufacturers.

The nature of the problems encountered can be presented by a discussion of a few typical cases. A good example is the equipment which is required for the operation of a Geiger-Müller counter when the actual number of pulses are to be recorded. The block diagram in Fig. 7 shows the essential components of this equipment. They may be arranged in separate units as indicated in the figure, or these components may all be assembled in a single unit. Although the trend is toward consolidation, there are commercial firms which manufacture scalers only, others which make mechanical registers, and still

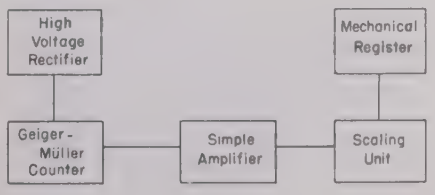


Fig. 7—Block diagram showing essential components of Geiger-Müller counting equipment.

Fig. 8—Geiger-counter scaler assembly with all components in a single chassis. *T*—electric timer; *V*—high voltage voltmeter; *M*—mechanical register; *S*—interpolation lamps of scaler.

others that make only electric timers. Also, there are a number of firms which make Geiger-Müller tubes but none of the accessory equipment. The nature of the tests required on each component will be given. A scaler in which all components are mounted in a single chassis is shown in Fig. 8.

We will start with the detector of radiation, the Geiger-Müller tube. These tubes fall into general classifications which differ in structural form, depending on whether they are designed to detect and record gamma rays or beta rays. Although a counter designed to detect beta rays is also sensitive to gamma rays, its efficiency for detecting beta rays is of the order of 50 to 100 times that for gamma rays, so that the effect of gamma rays is usually neglected. Gamma-ray counters are usually totally insensitive to beta rays of the usual energies. The basic principles on which these two types of counters operate are the same. Consequently, some tests are common to both types. One of the principal tests which is common to the two types is related to the characteristic curve of the Geiger-Müller counter, or plateau, and pulse height.

The plateau curve shows the pulse rate of a counter plotted against the voltage applied to the counter for a steady intensity of radiation on the counter. A typical plateau is shown in Fig. 9. The voltage T at which the counter begins to respond to the radiation is referred to as the starting voltage of the counter. Above this voltage the counting rate rises rapidly for a short interval of voltage, and then at A levels off in a gradually sloping line to a point B, where further increase in voltage produces another rapid rise in counting rate. The portion A to B of the curve is called the "plateau." The desirable features for this characteristic are that the value of T should remain fixed for a given counter; that the slope of the curve from A to B should be small (of the order of 1 or 2 per cent per hundred volts), and that the length of the plateau AB should extend over 200 to 300 volts. In addition, the counter is required to give a single clear-cut pulsef or each response to radiation. Multiple pulses are objectionable since they reveal that the counter is not functioning as a true Geiger-Müller counter, and because they either increase the dead time of the counter or falsify the measurements.

The dead time of a Geiger-Müller counter is the time after the production of a pulse during which the counter is in a state where it cannot produce a pulse, even when radiation has produced ions in the sensitive volume. This dead time is a direct result of the quenching mechanism of the counter which causes a discharge in the counter to cease abruptly, resulting in a definite electrical pulse. This time varies from about 10^{-3} to 10^{-4} second for current types of counters. It is desirable that the dead time be as short as is consistent with proper operation of the

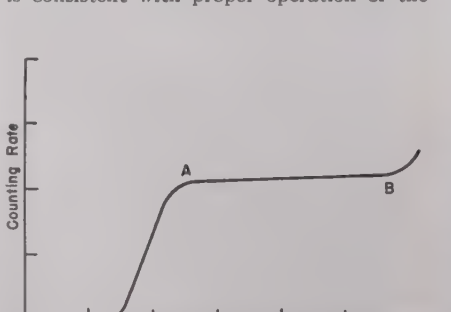


Fig. 9—Plateau of the Geiger-Müller counter. *T*—starting voltage; *A–B*—working region.

counter, since the length of the dead time termines the number of lost counts at a gi steady average counting rate. If the cour is to be operated at rates where the counts are a measurable percentage of total counts, the length of the dead time m be known to make a computed correctio the observations, in order to obtain the number of counts.

From considerations similar to those lined above, we arrive at the following t which are required of all types of Gei Müller counters for their use in quantita measurements.

1. *Determination of Background Coun Rate:* The actual background counting of a counter may be considered as made u three factors: (a) the intrinsic backgro rate; (b) response to cosmic rays; and (c) response to local distributed radioactiv The actual rate will depend on the condi under which the counter is used. Respons cosmic rays and local radioactivity can controlled to a considerable extent by shi ing. They are, therefore, under the con of the user to a certain degree. He has control over the intrinsic background r which is produced by radioactive radiat originating within the counter. This rate be controlled only at the time the count constructed by selecting materials with natural radioactivity, and by using car techniques to avoid the introduction of dioactive contamination into the coun The background rate controls the lower l of the net detectable counts from a sou which for many applications is import Therefore, the intrinsic background should be low.

The test for intrinsic background ra the obvious procedure of placing the cou inside a heavy lead shield to screen off radioactive radiations and to reduce the sponse to cosmic rays. In this way the lo background rate which can be expected der laboratory conditions is obtained. I intrinsic background is excessive, this reveals it immediately. It is obvious that intrinsic background depends on the are the interior surfaces of the counter. Beta counters with a tube of about 1 inch dia ter and $2\frac{1}{2}$ inches long may have an ove background when shielded with 2 inche lead, of from 0.3 to 0.5 counts per secon

2. *Determination of Plateaus for Ge Müller Counters:* The procedure for obt ing the plateau of any type of counter i sentially the same. The measurements made with the counter exposed to a stant source, which gives a convenient co ing rate. A scaler and mechanical registe used to record the counts. Before meas ments can be started, it is necessary tha high voltage be applied to the counter adjusted to the value at which the cou begins to count, called the starting vol This voltage is sometimes called the thres voltage, but it seems better to reserve latter term for the voltage at which pulses first become of uniform height i counter. The threshold thus defines the ginning of the true Geiger-Müller re The starting voltage must be applied to counter for a period of 15 to 20 minutes low electric charges on various insula parts to assume steady values. This is tr both the beta-ray and gamma-ray typ

ters. During this period the starting
age is subject to slow changes. When this
ing voltage has become constant, it is
to assume that static charges hav-
hed an equilibrium value.

The procedure for making the observa-
s consists in raising the high voltage by
roximately equal arbitrary steps, of the
r of 25 volts. At each voltage the time
ired to record a fixed number of counts
served. The number of counts depends
he statistical accuracy required. For a
dard deviation of 1 per cent, a total of
00 counts must be recorded for each
t. In recording the data for the plateau
essential to watch for a sudden increase
e counting rate carefully as the voltage
creased. Most counters in use today are
ily ruined if voltages above those cor-
onding to the plateau are applied. The
ting excessive rate quickly destroys the
nic vapor used for quenching.

The usefulness of a counter depends on
ength of the plateau and its slope. The
e may be defined as the percentage
ge in counting rate per 100 volts change
ne high voltage. A counter can be ex-
ed to have a slope of the order of 2 or 3
ent per hundred volts over a range of at
200 volts.

When plateaus must be plotted for large
bers of counters, it may be convenient
o this automatically. A device for the
ose has been described by Crits and
lin.[2] A rate meter with a recording mi-
nmeter is used to draw a curve of the
ting rate versus the time. The voltage
ied to the counter is automatically in-
sed by one step every 15 or 20 seconds
eans of an automatic timer. A rate me-
onnected to the counter actuates a re-
er which draws a curve consisting of a
s of steps. The steps are large at the
ning of the characteristic, small in the
n of the plateau, and larger above the
au. To protect the counter from exces-
voltages, a relay is arranged to discon-
e the operation when the rate meter ex-
s a predetermined reading. This equip-
t also could be arranged to proceed from
counter to another and draw the pla-
s for a considerable number of counters
out attention from the operator.

. *Effects of Changes of Temperature on
ters:* The starting voltage of most Gei-
Müller counters depends on the tempera-
This is particularly true of self-quench-
ounters which use an organic vapor as a
ching agent. A shift of the starting volt-
ends to move the whole plateau curve to
ight or left, with respect to the voltage
ed to the counter. If the temperature
t is large, a change of temperature of a
legrees during operation may throw the
ter entirely off of the plateau. In an
it will affect the counting rate to an ex-
dependent on the temperature coeffi-
, the temperature change, and the slope
e plateau. Therefore, it is important to
the effect of temperature on the count-
roperties of the counter.

or determination of the effects of tem-
ture, it is convenient to mount the
ter in a roomy box, well-insulated ther-

[2] J. Crits and G. C. Newlin, "Recording type
atic Geiger tube calibrator," U. S. Atomic En-
Commission Report 2025; 1948.

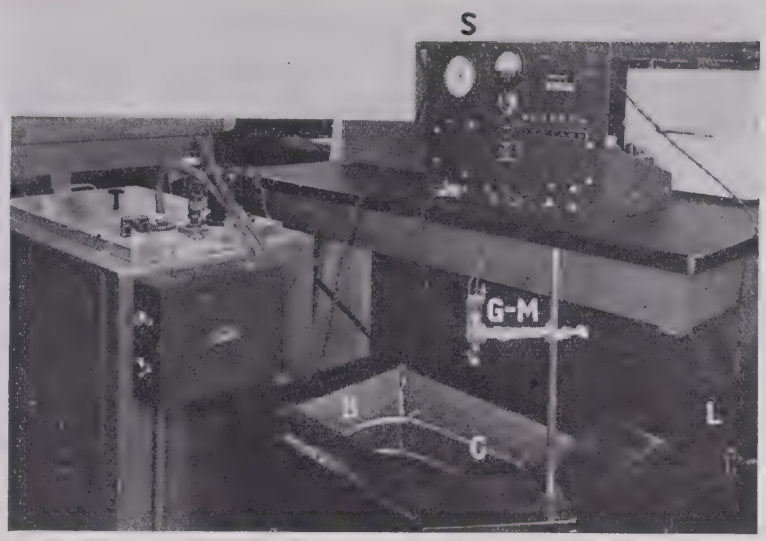


Fig. 10—Chamber with controlled temperature for determining the effect of temperature on Geiger-Müller counters. *T*—temperature control unit; *B*—temperature chamber; *C*—copper coil; *G–M*—Geiger-Müller counter ready to be lowered into chamber; *S*—scaler; *L*—lid of chamber.

mally. To control the temperature, a coil of copper tubing is mounted inside the box. The coil consists of a large number of turns mounted near the walls and extending over the greater part of the interior surface. The air inside the box is circulated rapidly by a fan, to aid in decreasing the time required for temperature equilibrium. Liquid at a controlled temperature is pumped through the copper coil. For most applications, tests conducted between 0°C and 30°C give adequate information. When operating at the lower temperatures it is essential to have a drying agent, such as silica gel, inside the thermal chamber to prevent condensation of water on the counter and leads. The temperature of the interior of the box is measured by several thermometers extending through the wall. No observations are made until these thermometers show a steady and uniform temperature. A photograph of this chamber is shown in Fig. 10.

To determine the temperature coefficient of the starting voltage, a constant source is mounted in a fixed position, chosen to give a counting rate of approximately 50 counts per second at the middle of the plateau. With the interior at the desired constant temperature, the data required for plotting the plateau are taken. This gives not only the effect of the temperature on the starting voltage, but reveals any effects on the lengths and slope of the plateau. With some filling mixtures, such effects are noticeable. For those counters where the principal effect of change of temperature is to alter the starting voltage, a temperature coefficient of 4 or 5 volts per °C is usually considered acceptable, although counters which have a lower coefficient have been encountered.

4. *Dead-Time Measurements:* As has been indicated, a knowledge of the dead time of a counter is often useful. Wherever a counter is operated at a counting rate such that the loss of counts due to dead time is appreciable, it is necessary to make a correction for this loss. Actually, the resolving time of the counter depends on the sensitivity of the equipment used to detect the pulses in the counter. The difference between the two is usually small, and the resolving time can never be less than the dead time. The relation between them is illustrated in Fig. 11.

The simplest observation of dead time is made according to the method developed by Stever,[3] using an oscilloscope provided with an external synchronization terminal and a driven sweep. The center wire is coupled through a capacitor of adequate voltage rating to the *y*-axis terminals of the oscilloscope. The amplified pulses from the counter are simultaneously applied to the external synchronization terminal. This results in the triggering of the sweep for each pulse. A source is placed near the counter to give between 100 and 200 counts per second. The pattern on the scope is that shown in Fig. 11.

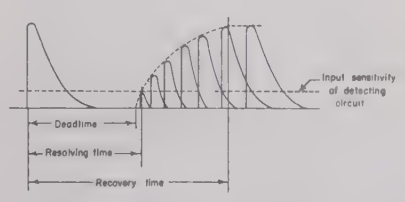


Fig. 11—Diagram showing recovery of a Geiger-Müller counter, subsequent to an initial pulse. In the interval labeled "dead time" no other pulse appears. The recovery time is determined in part by the size of the pulse which can be detected by the indicating circuit. The recovery time is determined by the time required for the pulses to regain their original size.

[3] H. G. Stever, "The discharge mechanism of fast G-M counters from the deadtime experiment," *Phys. Rev.*, vol. 61, pp. 38–52; January, 1942.

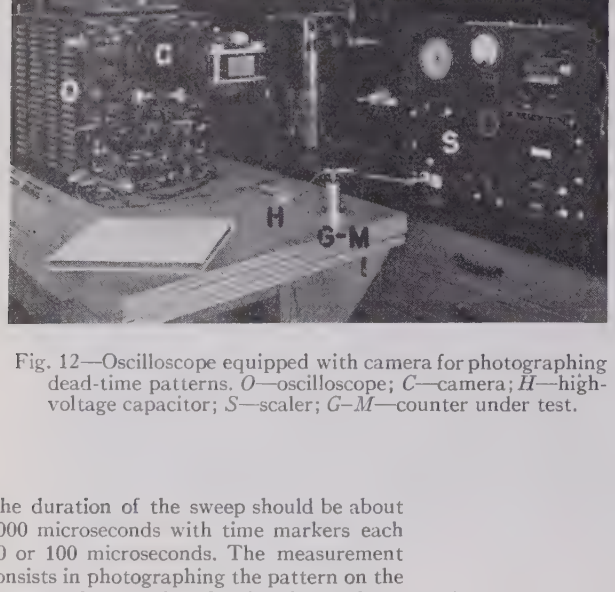


Fig. 12—Oscilloscope equipped with camera for photographing dead-time patterns. *O*—oscilloscope; *C*—camera; *H*—high-voltage capacitor; *S*—scaler; *G–M*—counter under test.

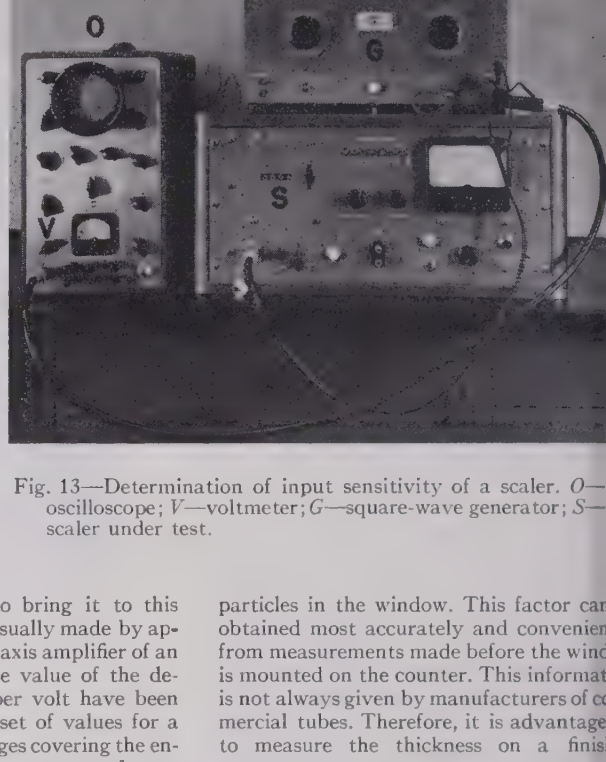


Fig. 13—Determination of input sensitivity of a scaler. *O*—oscilloscope; *V*—voltmeter; *G*—square-wave generator; *S*—scaler under test.

The duration of the sweep should be about 1000 microseconds with time markers each 10 or 100 microseconds. The measurement consists in photographing the pattern on the screen and measuring the time intervals on the negative. Fig. 12 shows equipment for photographing this pattern.

For more accurate measurements, a method of electronic gating may be used. This refinement is not always necessary in the routine tests of counters, but it is helpful in investigations of factors which affect the dead time. By detection of small changes in dead time, it reveals the direction in which improvement has been achieved. This information may result in further modifications which give still greater improvements. A suitable method for using electronic gating is described in a forthcoming issue of the *Journal of Research of the National Bureau of Standards.*[4]

5. *Measurement of Pulse Size:* The sizes of the pulse in volts from Geiger-Müller counters vary from counter to counter, depending on the size of the electrodes, the filling gas, and other constructional features.

It is desirable to know the pulse size of a counter at some representative voltage, since this determines whether the counter can be used with a given amplifying and detecting equipment. The difference in voltage between the operating and threshold voltages is defined as the overvoltage. The size of the pulses in a given counter increases with overvoltage. Therefore, some definite part of the plateau must be selected in comparing pulse sizes in counters. A knowledge of this pulse size is needed to determine whether a counter can be used with a given amplifying and detecting equipment. In addition, the pulse size is a function of the capacitance across the counter. Therefore, in making comparative measurements it is essential that observations be made for a definite fixed value of this capacitance, which includes a number of elements in an actual experimental arrangement. These include the cable and coupling capacitors and those in the instrument used for observations. It is usual to select either 50 or 100 $\mu\mu$f for this capacitance, and adjust the connecting system to bring it to this value. Observations are usually made by applying the pulses to the *y*-axis amplifier of an oscilloscope for which the value of the deflections in millimeters per volt have been ascertained. A complete set of values for a number of arbitrary voltages covering the entire plateau is taken. For purposes of comparison, the size of the pulse at the center of the plateau is a convenient parameter. The equipment for these measurements is shown in Fig. 13.

6. *Photosensitivity:* Finally, it should be mentioned that it is desirable to ascertain whether counters are photosensitive. Such a test should be applied to all counters constructed in a way which permits light to strike the interior electrodes. When a counter is photosensitive, its response to light is usually manyfold greater than to radioactive radiations under usual conditions of operation. However, this is not always true and photosensitivity is not always recognized immediately. It may then lead to erroneous conclusions regarding the functioning of the counter or its associated equipment, and is always a possible source of error in measurements. This defect of counters is readily detected by enclosing the counter in a light-tight shield provided with a small aperture through which light may be directed. A counter which is photosensitive must either be used in a light-tight enclosure or completely coated with opaque varnish. Since photosensitivity occasionally develops with use, it is a good precaution to screen all counters from light unless they are constructed in such a way that light does not reach the interior.

The above tests are appropriate for any type of Geiger-Müller counter. There are a few tests, in addition, that may be applied to beta-ray counters only, and, similarly, one or two which relate to gamma-ray counters. For beta-ray counters, perhaps the most important is a measurement of the thickness of the window. In quantitative measurement with beta-ray counters it is important to know the thickness of the window to permit application of corrections for absorption of beta particles in the window. This factor can obtained most accurately and convenien from measurements made before the windo is mounted on the counter. This informat is not always given by manufacturers of co mercial tubes. Therefore, it is advantage to measure the thickness on a finis counter.

One method of measuring the thickn of mica windows on end-window coun which has been applied with good results volves the use of polonium alpha partic This method can be applied to windows in range of 1 to 4 mg/cm², which includes th used on most commercial counters. The p cedure is to mount the counter in a rig fixed position, with the window end do Below this a polonium source is mounte such a way that its distance from the w dow can be determined accurately. source can be moved vertically by means micrometer screw with a vernier scale c brated in millimeters. A photograph of arrangement is shown in Fig. 14. The cou is operated in the proportional region, wh it only detects alpha particles, and pu are recorded by use of a linear amplifier scaler. The observations consist in plot the counting rate against the distance of source from the window. A typical curv shown in Fig. 15. The steep part of curve is extrapolated to the *x*-axis, which g the extrapolated range of the polonium al particles minus the window thickness. extrapolated range of polonium alpha ticles is 3.897 cm at 15°C, 76 cm press The observations are corrected for temp ture and pressure and subtracted from above value for the range which gives thickness of the window in centimeter air. This value is converted to mg/cm give the thickness in the units comm used. This method has been applied to dows of known thickness and gives excel agreement.

Optical methods can be used for measu thin, transparent films, as has been pos

[4] L. Costrell, "Accurate determination of the dead-time and recovery characteristics of Geiger Muller counters," *Jour. Res. Nat. Bur. Stand.*, vol. 42, pp. 241–251; March, 1949.

[5] M. G. Halloway and M. S. Livingston, "R and specific ionization of alpha particles," *Phys.* vol. 54, pp. 18–37; January, 1938.

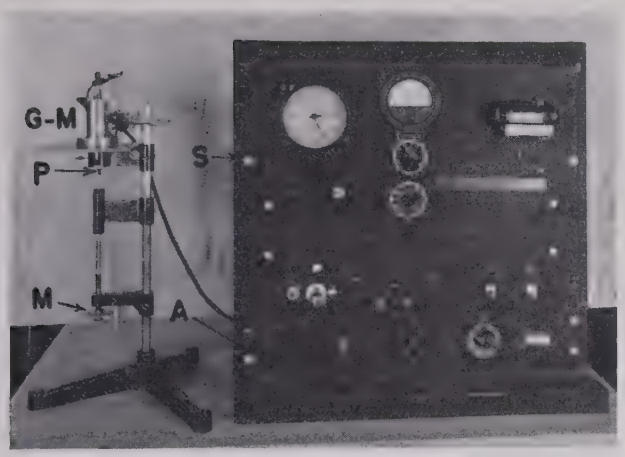
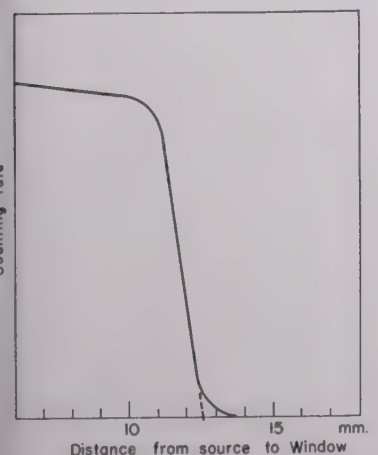


Fig. 14—Equipments for measuring thickness of mica windows in counters using α particles from polonium. *G–M*—mica window counter; *P*—polonium source; *M*—micrometer screw; *A*—linear amplifier; *S*—scaler.

by Wood.[6] Recently, Brown and Willghby[7] described a practical adaptation of procedure. The window is illuminated by ordinary incandescent lamp at an angle of dence of 10°. The reflected light, when wed by a small direct-vision spectroscope, ws a continuous spectrum crossed by inference fringes. The number of fringes ween two lines in a comparison spectrum counted. The comparison spectrum can obtained from a fluorescent light, and it onvenient to take the 577-mμ yellow line the 436-mμ blue line of mercury as refere points. An experimental curve gives the tion between the number of milligrams square centimeter and the number of ges between the reference lines. W veight in mg/cm^2; ρ is measured density the mica = 2.85; μ is the index of refrac- = 1.58; γ is the angle of incidence;



. 15.—Graph showing the number of alpha particles detected as a function of the distance of the polonium source from the window of the counter.

[6] R. W. Wood, "Physical Optics," p. 192.
[7] F. W. Brown, III, and A. B. Willoughby, "Optimethod of determining thickness of Geiger tube dows," *Rev. Sci. Instr.*, vol. 19, pp. 820–821; Noiber, 1948.

$\lambda_1 = 577$ mμ; $\lambda_2 = 436$ mμ; and n is the number of fringes. The formula for the curve reduces to

$$W = 0.163n \frac{\text{mg}}{\text{cm}^2}$$

for an angle of incidence of 10°, where W = thickness in mg/cm^2, and n is the number of fringes.

For gamma-ray counters it is useful to know the relative efficiency of a counter for gamma radiation of various energies. This can be determined by coincidence measurements between the pulses from the gamma-ray counter and a beta-ray counter when both are exposed to the same preparation of a radioisotope emitting both beta and gamma rays, provided the energy of the gamma rays is known and the disintegration scheme is reasonably simple. Strictly, this method requires that the isotope emit a single gamma ray, and relatively few radioisotopes have such a simple disintegration scheme. However, it is possible to find several which emit two or three gamma rays with energies near enough together to give a reasonable average for the energy of the group. Among these are Na^{23}, Na^{24}, Co^{60}, and Au^{198}. The procedure consists in observing the counting rates of the two counters separately and also the coincidences between them, using a Rossi-type coincidence circuit, illustrated in Fig. 2. The efficiency of the gamma-ray counter in the particular arrangement used is given from the following considerations:

Let A = counting rate of gamma counters
B = counting rate of beta counters
C = coincidence rate
N = disintegration rate of source, which need not be known.

If E_γ is the efficiency of the gamma counter and E_β that of the beta counter,

$$A = E_\gamma N, \quad B = E_\beta N \quad \text{and} \quad C = E_\gamma E_\beta N.$$

Therefore,

$$E_\gamma = \frac{C}{E_\beta N} = \frac{C}{B},$$

or the efficiency of the gamma counter is the ratio of the coincidence rate to the counting rate of the beta-ray counter. This experiment can be repeated, maintaining a fixed geometrical arrangement of counters and source, using radioisotopes emitting gamma rays of various energies to give the desired information regarding the variation of sensitivity of the counter with the energy of the gamma rays.

7. *Test of Scalers:* A scaler for use with counters usually consists of a number of stages, each containing two electron tubes, connected in a circuit so that only one can be in the conducting state. This arrangement permits each stage to transmit one-half the pulses which enter it. Therefore, the scaling ratio is 2^n where n is the number of stages. Decade scalers consist of a modification of the scale-of-two connections in stages so that every tenth pulse is transmitted by a stage. Although there is usually some type of electron-tube amplification between the counter and the scaler, in practice this amplifier is included with the scaler assembly. This permits operation of the counter by a direct connection to the scaler unit. Tests of the scaler unit therefore automatically include the amplifier.

Scalers are usually tested to determine the minimum pulse voltage to which the scaler will respond as well as the maximum pulse voltage at which it fails to operate satisfactorily. The minimum pulse on which a scaler will operate satisfactorily may be determined by use of a square-wave generator of variable amplitude synchronized with the 60-cps power line. The square pulses are passed through a differentiating network to convert them into pulses with a steep rise and an exponential decay with a time constant of approximately 50 microseconds. The alternator of the square-wave generator is set for a minimum voltage output, and the pulse size increased gradually until the scaler responds accurately to the 60-cps rate. This pulse size is measured on a cathode-ray oscilloscope, and is the minimum input sensitivity of the scaler. The pulses may then be increased in voltage until the scaler begins to show erratic operation, which gives the maximum voltage of pulse for the scaler.

The input sensitivity may also be checked under actual operating conditions, using pulses from a Geiger-Müller counter. With a counter connected to the scaler, the voltage on the counter is set at approximately 100 volts above the threshold. Assuming that the scaler then functions correctly, which can be checked by observing counting rates with sources of known ratios of intensity, the voltage on the counter is reduced until the scaler begins to fail to respond accurately. If this occurs while the counter is still operating in the Geiger-Müller range, the pulses will still be of uniform height. This height can be measured with a cathode-ray oscilloscope, as before.

The scaling accuracy is most simply checked by using a generator synchronized with the 60-cps power line and timed by a electric clock driven by the same supply. The pulse size is set at about 1½ times the minimum input sensitivity, and the clock and scaler started simultaneously. At the end of an hour the elapsed time is compared with the recorded counts. If it is desired to

check the accuracy of scaling at other frequencies, a stable oscillator which can be set at a number of frequencies is substituted for the 60-cps synchronization, and the observations are repeated. Sometimes a scaler is found which will scale correctly over a small range of input pulses, but will fail for larger pulses. It is usual, therefore, to increase the pulse size to several times the lower limit of input sensitivity in steps, and determine the scaling accuracy at each step.

In the test of scalers it is important to check all control switches, particularly the "start and stop" switch, to ascertain whether the scaler starts reliably when this switch is thrown to the "start" position and whether this operation introduces spurious counts into the scaling stages. Commercial scalers have been encountered which do not always start scaling when this switch is thrown to the starting position. The introduction of spurious counts by the switch is also a defect encountered occasionally.

It is desirable to check the operation of scalers over extended periods of time, since failures due to overloading components become evident in this way. Continuous operation for twenty-four hours followed by the usual tests reveals defects of this type, and also reveal whether sufficient ventilation has been provided to prevent overheating.

Most scalers operate satisfactorily over a range of from 90 to 130 volts on the power-line input. In a few instances scalers have been encountered which become erratic if the line voltage fluctuates ± 10 volts. To detect whether this characteristic is present, scalers are tested with the supply voltage at 90 volts and at 130 volts, as well as at 110 volts.

The high-voltage rectifier is tested for regulation. Since counters usually have plateaus with slopes of the order of 5 per cent per 100 volts, it is important that the high voltage remain stable within better than 1 per cent for the fluctuations in power-line voltages ordinarily encountered. A rough test of the stability can be made by using a voltmeter of appropriate range and varying the supply voltage over a range of 90 to 130 volts. If changes in the voltmeter readings are readily discernible, this usually means that the high voltage is insufficiently regulated. Where it is desirable to determine the actual degree of regulation, a null-type difference voltmeter circuit is used. This permits measurements of changes of less than 1 volt in the high voltage.

The above tests are also applied to ascertain the stability of the high voltage with time for a constant regulated supply voltage. To obtain a satisfactory test of this kind, it is necessary to make observations on the high voltage over short intervals of time, say every minute, for 10 minutes, and over longer intervals of the order of every 10 minutes for an hour. Slow changes are detected by observations every hour over a period of several hours.

If the high-voltage rectifier is supplied with a voltmeter, which is desirable, it is necessary to check the accuracy of this voltmeter. This meter, to be of real use, must read standard volts to within a few per cent. It is usually relied upon to adjust the voltage for operation of a new counter. If it is off by more than a few per cent, there is danger of applying an excessive voltage to the counter before the operating range for the counter has been ascertained in terms of the inaccurate meter. It is equally important that the errors be uniformly distributed over the scale. The voltmeter can be checked at a number of readings by use of a calibrated voltmeter of high resistance connected in parallel with it.

The only component of the scaling-circuit assembly remaining to be discussed is the mechanical register. Fortunately, these devices are ruggedly built, and are not required to operate at very high recording rates. When high counting rates are used, additional scaling stages are usually introduced to relieve the register of excessive rates. It is rare that a register of the decade type fails to operate properly, and little attention is given to them in testing unless obvious defects become apparent.

8. *Testing of Rate Meters:* Equipment for the operation of Geiger-Müller counters using the principle of the rate meter for visual indication is less commonly used for quantitative measurements than is the scaler type of instrument. This situation arises from the fact that the indicating meter is subject to fluctuations inherent in the emission of radioactive radiations, and is therefore difficult to read accurately. These fluctuations become pronounced in the more sensitive models with a short time constant to give more nearly instantaneous readings of intensity. Rate meters are usually employed in counter units for survey purposes where high accuracy is not expected. They also lend themselves readily to continuous recording on moving-chart recorders. Rather simple tests are required on these instruments, such as determinations of the stability of the high-voltage supply, variations of meter indication with power-line voltage, and a test of the linearity of the meter scale with the intensity of the radiation incident on the counter. Usually, commercial models are supplied with a counter tube, so that tests are made on the complete assembly. Occasionally, rate-meter dials are calibrated in counts per unit of time. The accuracy of these scales can be checked by use of a stable oscillator with adjustable frequency to produce suitable pulses for activating the circuit.

9. *Health Survey Meters:* A considerable number of instruments manufactured for the measurement of radioactivity consist of devices using ionization chambers to measure gamma radiation in roentgens. The visual indication is achieved either by quartz-fiber electroscopes, or by electronic amplification of the ionization current. The accuracy required of these meters is of the order of ± 15 per cent.

The most important test of survey meters consists in checking the accuracy of the scale of the instrument, which usually is marked to read milliroentgens per hour. Those using quartz fibers require observation of the rate of deflection of the fiber. In some cases a timer, such as a flashing light, is built into the instrument.

The calibration of these survey meters is conveniently checked by use of the gamma radiation from standardized preparations of radium. One milligram of radium enclosed in 0.5 mm of platinum gives, with sufficient approximation, 0.85 mr per hour at a dista of 1 meter. Therefore, preparations of dium of the order of 10 to 100 mg are c venient in checking the calibration of usual survey meter. The inverse square l can be assumed to hold for distances of order of 25 to 200 cm, if care is taken to duce the effects of scattering of gamma rac tion by near-by objects. Measurements distance of the source are made from the c ter of the ionization chamber. More accur tests are obtained if a series of radium pre rations are available, so that the indicati of the instrument at differeng intensities the radiation are made in a fixed geomet

The variation of the sensitivity w temperature is made at two temperatur about 20°C apart. Temperatures in the ne borhood of 0°C and 20°C are usually chos for this test.

Survey meters should give readings terms of roentgens, which means that th should indicate 1 roentgen for that intensi of radiation which produces the amount ionization in air required by the definition the roentgen. Unless the ionization chamb has been properly designed to produce t same relative ionization as is obtained in air-wall chamber for gamma radiations various energies, the chamber will not gi true readings for all energies of gamma ra Such chambers are sometimes called "e ergy-dependent." The degree of deviati with energy in these chambers can be check by comparing the readings with those of instrument which has been checked over range of energies when both are exposed su cessively to radiations of differing energi In practice, this is most conveniently acco plished by using as sources a series of radi isotopes which emit gamma rays of differe energy. This tests the instrument under co ditions approximately those under which will be used.

Some health survey meters are of the i tegrating type; that is, they indicate the to amount of radiation in milliroentgens which they have been exposed. The most i portant of these is the pocket chamber. Th is a small ionization chamber provided wi best quality of electrical insulation. In u the central electrode is charged to a defin potential, and the chamber is issued to an i dividual whose work exposes him to gam radiation. At the end of the day the chamb is returned to the laboratory, and the resi ual potential measured. The charging d vice contains a quartz-fiber electrosco which measures the initial charging pote tial, and is also used to measure the residu potential. The scale of this instrument is c ibrated in roentgens.

Pocket chambers are tested for natu leakage of charge due to imperfect insulatio In a good chamber a negligible fraction the initial charge will leak off in twenty-fo hours when protected from exposure to diation. Tests are also made of the loss charge due to vibration or shock. A go chamber may be dropped on the floor fr height of several feet and show no change potential. Finally, the accuracy of the ca bration is checked by comparing the rea ings obtained from the charging electrome with the radiation computed from exposu of measured duration at various distan from standardized preparations of radiu

The Columbia Long-Playing Microgroove Recording System*

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Summary—The Columbia LP (long-playing) microgroove re-ing system was developed to fill the need for music reproduction ch would avoid interruptions not intended by the composer, and ch would be of excellent quality at a reasonable cost. This all-ortant factor of cost and the public's familiarity with the handling honograph records made it desirable to solve the task on the s of records, rather than tape or wire.

Standard 78-rpm records were originally designed to generate id mechanically by direct transfer of energy from the groove of record to the vibrating diaphragm. Because the entire acoustical gy had to be extracted from the grooves, these had to be quite ed, and remained so up until now.

The new Columbia recording system was an inevitable outcome he use of electrical amplification between the groove and the speaker. Today, practically no mechanical energy needs to be acted from the groove, and thus, for the first time, it has been ible to develop much finer grooves, permitting longer playing and distortion-free reproduction.

Introduction

THE COLUMBIA LP[1] (long-playing) microgroove recording system was developed to provide uninterrupted music reproduction of better quality a reasonable cost. This all-important factor of cost, ether with the public's familiarity with the handling phonograph records, made it desirable to solve the blem on the basis of disk records, rather than by tape wire.

Standard 78-rpm records were originally designed to erate sound mechanically by direct transfer of energy m the groove of the record to the vibrating dia-agm. Because the entire acoustical energy had to be racted from the grooves, these had to be quite ged, and remained so until now. The weight of the phragm and armature, together with the acoustical roducer, resulted in high vertical needle forces, re-ring large stylus diameters. The latter made neces-y the use of the high-speed turntable (78 rpm) in er to allow for a sufficient frequency range and a imum of distortion. Today, practically no mechani- energy need be extracted from the groove, thus al-ing a far greater latitude in the design of an im-ved system.

The fundamental principles of the Columbia LP rec-s are the slow rotational speed of 33⅓ rpm, represent- a time-saving factor of roughly 2.3; and finer oves, up to 300 per inch.

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† Columbia Broadcasting System, Inc., New York, N. Y.
‡ Columbia Records, Inc., New York, N. Y.
[1] Trade-mark of Columbia Records, Inc., New York, N. Y.

Playing Time

It was found that the average playing time for classical works is approximately 36 minutes. Nevertheless, standards were developed to accommodate as much as 50 minutes of playing time on a 12-inch record.

The following analysis establishes the optimum relationship between the inside and outside groove diameters for maximum playing time.

The variables used are:

d = inside groove diameter in inches
D = outside groove diameter in inches
R = revolutions per minute
N = total number of grooves
T = playing time in minutes.

The constants used are:

n = grooves per inch
$V_{\min}$ = minimum permissible linear groove velocity in inches per second.

The playing time T is equal to the total number of grooves N divided by the revolutions per minute R of the record.

$$T = \frac{N}{R}. \tag{1}$$

N is a function of the number of grooves per inch and of the radial distance utilized:

$$N = n\left(\frac{D - d}{2}\right). \tag{2}$$

R is determined by the minimum recorded radius and the minimum permissible linear groove velocity:

$$R = \frac{60V_{\min}}{\pi d}. \tag{3}$$

Substituting (2) and (3) in (1),

$$T = \frac{n\left(\frac{D - d}{2}\right)}{\frac{60V_{\min}}{\pi d}} = \frac{\pi n}{120V_{\min}}(Dd - d^2). \tag{4}$$

Maximizing T with respect to d,

$$\frac{\delta T}{\delta d} = \frac{\pi n}{120V_{\min}}(D - 2d) = 0$$

and

$$d = \frac{D}{2} \cdot \tag{5}$$

In other words, maximum playing time for a given inner-groove linear velocity is obtained when the inner diameter of the record equals one-half the outer diameter.

From (4), with $d = D/2$

$$T = \frac{\pi n D^2}{480 V_{\min}}$$

and

$$n = \frac{480 V_{\min} T}{\pi D^2} \cdot \tag{6}$$

Thus, when $T = 20$ minutes, $V_{\min} = 10$ inches per second and $D = 11.5$ inches; the number of grooves per inch n will be 230. As the nearest practical value, 224 grooves per inch are used for all records up to 20 minutes per side, or 40 minutes total playing time. This requires an inner diameter of 5.5 inches, with a corresponding velocity of 9.6 inches per second. On record sides requiring more than 20 minutes of playing time, 260 grooves per inch are used.

Tracing Distortion

Increasing the length of playing time by an appreciable factor was only one of the goals to be accomplished. With the advent of FM and professional tape recorders, improved performance over standard 78-rpm and even transcription records became a necessity.

In the following, an analysis will be made of the performance of Columbia's LP records in relation to commercial 78-rpm records, and also to transcription-type recording. The chief factors considered will be those which are influenced by the linear speed and by the amplitude excursions of the reproducing stylus. It has been found in disk recording in general that the limitations imposed by cutter performance, such as clearance angle of cutting stylus, acceleration of the stylus, etc., are usually less limiting than the tracing distortion, which is a function of the minimum radius of curvature of the traced waves and the effective radius of the reproducing stylus.[2]

First, the minimum radius of curvature corresponding to maximum deviation amplitude for various types of records (78-rpm, transcription, and LP) will be determined; at the same time, it will be postulated that the radius of curvature is not to be less than the effective radius of the reproducing stylus. This does not imply that such a condition ensures distortion-free reproduction, since, even under this condition, the center of th reproducing stylus does not trace a sine wave. Disto tion increases very rapidly when the effective styl radius exceeds the radius of curvature.[2] The purpose this analysis, however, is to show the relative perforn ances of 78-rpm, transcription, and LP records, rath than to define certain requirements already well-know in the recording field.

Arbitrarily, the condition where the two rad (namely, minimum radius of curvature of wave and e fective radius of reproducing stylus) are equal, will b named the limiting condition, and the correspondin frequency will be termed the limiting frequency.

In general, the radius of curvature for a functio $y = f(x)$ can be expressed as follows:

$$r = \frac{\left[1 + \left(\frac{dy}{dx}\right)^2\right]^{3/2}}{\frac{d^2y}{dx^2}} \cdot \tag{7}$$

Applying this to a sine wave, $y = D \sin 2\pi x/\lambda$, it can b shown that r has a minimum for $x = \lambda/4$. Thus,

$$r_{\min} = \frac{\lambda^2}{4\pi^2 D} \cdot \tag{8}$$

The wavelength λ as traced on the record is equal to th linear velocity V of the groove at that point, divided b the frequency f of the energy delivered to the cuttin stylus. Therefore, the frequency at which the minimun radius of curvature will equal the effective reproducing stylus tip radius for the deviation D can be called limit ing frequency, and will be

$$f_l = \frac{V}{2\pi\sqrt{r_{\text{eff}} D}} \cdot \tag{9}$$

Actual values of f_l can now be determined for 78-rpr records, transcription disks, and LP records for th maximum possible deviation ($D_{\max}$) when using th innermost grooves. The critical frequency values are c greatest interest for those grooves nearest the center be cause of their low linear velocity.

For 78-rpm records at a 4-inch diameter, the linea velocity $V = 16.3$ inches per second. The tip radius i usually 3 mils, and the effective radius $r_{\text{eff}} = 0.003$ co $\alpha/2$, where α is the total groove angle. Thus, sinc $\alpha = 90°$, $r_{\text{eff}} = 0.0021$ inch. The peak displacemen $D_{\max} = 0.002$ inch; substituting these values in (9),

$$f_{l(78)} = \frac{16.3}{2\pi\sqrt{(0.0021)\cdot(0.002)}} = 1{,}270 \text{ cps.} \tag{10}$$

The limiting frequency for transcription records is

$$f_{l(TR)} = \frac{14}{2\pi\sqrt{(0.0017)\cdot(0.0011)}} = 1{,}620 \text{ cps.} \tag{11}$$

[2] W. S. Bachman, "Phonograph dynamics," *Electronic Ind.*, vol. 4, pp. 86–89; July, 1945.

inally, the innermost grooves of LP records have a ır velocity of the order of 9.6 inches per second; reproducing-stylus tip radius is 0.001 inch, and the tive radius equals 0.0007 inch. The maximum ve deviation is 0.0009 inch. From these values, it ws that

$$f_{l(LP)} = \frac{9.6}{2\pi\sqrt{(0.0007)\cdot(0.0009)}} = 1{,}940 \text{ cps.} \quad (12)$$

s, the limiting frequency for LP records is higher ı for either transcription records or standard 78- records.

Vith regard to these frequency values, an obvious clusion would be that none of the three recording ems under scrutiny is capable of reproducing fre- ıcies above the values corresponding to the limiting uency; in the case of the LP record, this would be 0 cycles. It is necessary to keep in mind, however, the various values for f were established using the imum groove displacement D_{max}.

ınce second-harmonic distortion is automatically inated by virtue of the complete symmetry of the ortion produced in the traced waves, another way of paring tracing distortion between 78-rpm commer- records, transcription records, and the LP micro- ve records is to utilize an expression for the rms ıe of third-harmonic lateral tracing distortion[3] T.

$$T = \frac{6\pi^4 \cdot r^2 \cdot D^2 \cdot f^4}{V^4} \cos^2 \frac{\alpha}{2}, \quad (13)$$

re

r = tip radius of the reproducing stylus in inches
D = peak amplitude of the recorded wave in inches
f = frequency of the recorded wave in cps
V = linear groove velocity in inches per second
α = groove angle.

ınce the purpose is to determine the relative distor- of various disk-recording systems, a graph has been ared (Fig. 1) in which, based on (13), the distortion LP records, commercial transcription records, the ntly announced RCA 7-inch record, and conven- al 78-rpm records has been determined in relation he tracing distortion on the inside of the standard pm disk. The ordinate gives the inverse of this dis- ion ratio, so that, the higher the ordinate value, the er the theoretical quality of the record. For in- ce, the LP record on the outside (zero minutes play- time) shows 85 times less distortion than the inside he conventional 78-rpm record. After 21 minutes ing time, the LP record shows roughly 6.2 times less ortion. The RCA 7-inch record shows 7.4 times less ortion. With respect to actual demonstrable differ- s among these various types of records, it can be seen that the tracing distortion on the inside of commercial transcription records (after 15 minutes playing) is roughly one-third of that found on the inside of the 78-rpm record. It is questionable whether additional reduction in tracing distortion can be demonstrated with anything short of professional laboratory equipment when using regular program material.

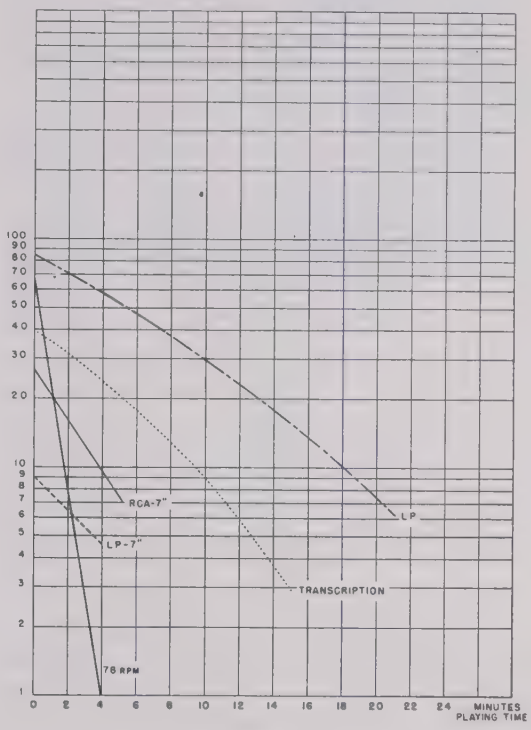


Fig. 1—Performance characteristics of various types of records.

Equation (9) may be written with $r_{min} = r_{eff}$.

$$D_{usable} = \frac{V^2}{4\pi^2 f^2 r_{eff}}. \quad (14)$$

This corresponds to the usable deviation that may be employed in recording a frequency f while still retaining the original condition that the stylus radius does not exceed the minimum radius of curvature of the recorded wave. From (14),

$$\frac{D_{usable}}{D_{max}} = \frac{\dfrac{V^2}{4\pi^2 f^2 r_{eff}}}{\dfrac{V^2}{4\pi^2 f_l^2 r_{eff}}} = \left(\frac{f_l}{f}\right)^2. \quad (15)$$

In Fig. 2, D_{usable}/D_{max} is plotted in percentage versus frequency for 78-rpm records, for transcription records, and for LP records. The ordinate indicates percentage usable maximum groove displacement where 100 per cent equals the maximum possible values used in (10),

W. D. Lewis and F. V. Hunt, "A theory of tracing distortion in d reproduction from phonograph records," *Jour. Acous. Soc.* ., vol. 12, p. 353, 1941.

(11) and (12). Thus, 100 per cent modulation equals 0.002 inch (10) for 78-rpm records; 0.0011 inch (11) for transcription records; and 0.0009 inch (12) for LP records. As indicated previously, these conditions are those prevailing at the lowest linear velocity (inside grooves) for each particular type of record.

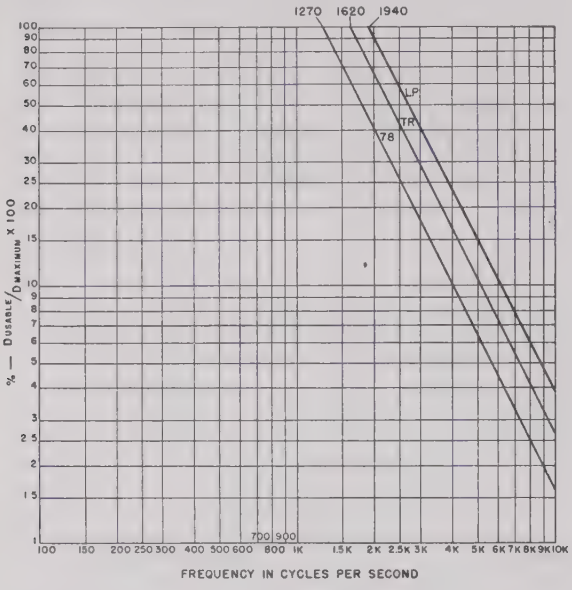


Fig. 2—Percentage maximum deviation versus frequency.

The LP recording characteristic (velocity versus frequency) is shown by curve 1 in Fig. 3. Curve 2 in Fig. 3 shows the NAB velocity recording characteristic, which deviates from the LP characteristic in the low bass por-

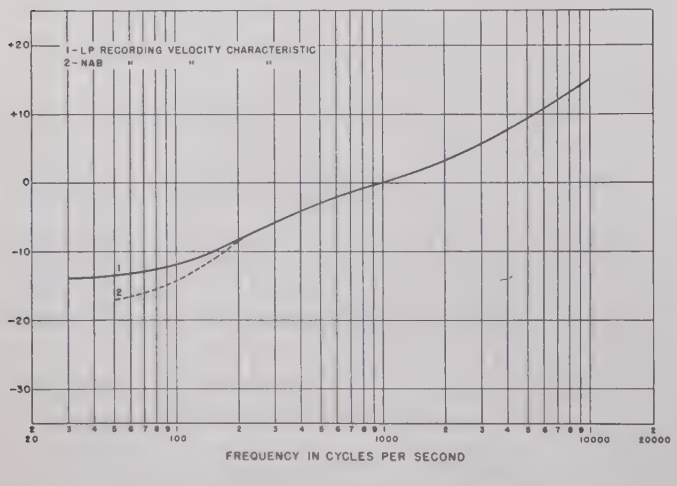


Fig. 3—LP and NAB recording characteristics.

tion only. The purpose for the LP bass lift is to reduce rumble and hum pickup, which would otherwise be more pronounced with the smaller deviations.

Hand in hand with the LP record, suitable reproducing equipment has been developed in the Columbia Laboratories. Existing pickups, such as used for rpm records, obviously could not be employed. Fi as has been mentioned previously, the tip radius LP records should be 0.001 inch (± 10 per cent). S ond, the weight requirements are quite different. I tensive tests have shown that, for a stylus of 0.0 inch radius, as used with the LP records, a downw stylus force not exceeding 6 grams is desirable. LP stylus employing a 6-gram force exerts no m pressure than the average 78-rpm pickup.

The necessity of developing a single cartridge capa of tracking both LP and 78-rpm records with only grams stylus force was also realized. As a result, t cartridges with a new nonresonant type of arm w developed; a single cartridge playing LP records on and another cartridge having two styli and a sin crystal capable of playing either LP or 78-rpm recor Both cartridges track with a downward force of only to 6 grams and have approximately 0.5 volt rms outp These cartridges, with an associated filter, have a su stantially flat response over the entire recorded f quency range.

Since the included groove angle of both LP and 7 rpm records is approximately 90 degrees, the vertic force F_v tending to force the stylus out of the groo is equal to the total lateral force on the stylus F_L. Th force has a component F_s which overcomes the stiffne of the stylus and tends to force the stylus away from i position of equilibrium, and an opposing component F which accelerates the stylus toward its position equilibrium. At low frequencies, where most trackin problems were encountered, F_a, which decreases wi the square of the recorded frequency, becomes negli ble compared to F_s, and can, therefore, usually be ne lected.

Then,

$$F_v = F_L = F_s \text{ (at low frequencies).} \quad (1$$

Now

$$F_s = \frac{D_{\max}}{C} \quad (1$$

where C is the compliance measured at the point of t stylus.

From (16) and (17),

$$F_v = \frac{D_{\max}}{C} \quad \text{or} \quad C = \frac{D_{\max}}{F_v}. \quad (1$$

Substituting the $D_{\max}$ and F_v used in the LP system (18),

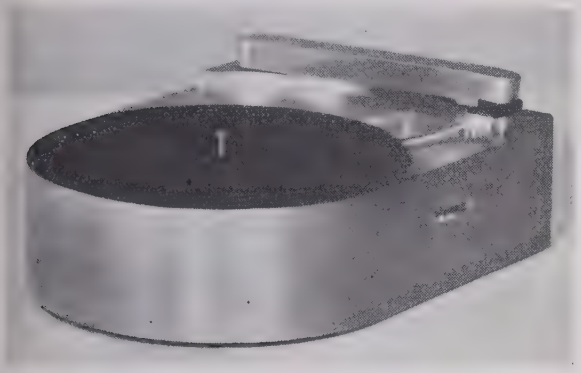
$$C = \frac{(0.0009)\cdot(2.54)}{(6)\cdot(980)} = (0.39)\cdot 10^{-6} \text{ cm/dyne.} \quad (1$$

Substituting the corresponding values for the 78-r side of the dual cartridge,

$$C = \frac{(0.002)\cdot(2.54)}{(6)\cdot(980)} = (0.87)\cdot 10^{-6} \text{ cm/dyne.} \quad (20)$$

This higher value of compliance was obtained by ıploying a stylus wire of a smaller diameter. Actual mpliance values in these cartridges are more than uble the values given above, which represent the west ones required theoretically.

A number of laboratory player attachments were ilt using the newly developed pickups and employing ndard two-pole motors, modified to give wow- and mble-free reproduction. The players and the records re field-tested in homes over a considerable period time. Fig. 4 shows the LP attachment designed by



ig. 4—Columbia's experimental dual cartridge-dual speed player.

olumbia to play both 78-rpm and LP records using e nonresonant arm and dual cartridge described previously. Fig. 5 shows the Columbia player attachment engineered by Philco.

Fig. 5—Philco LP record player.

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Direct Voltage Performance Test for Capacitor Paper*

HAROLD A. SAUER† AND DAVID A. McLEAN†

Summary—Performance of capacitors on accelerated life test ay vary over a wide range depending upon the capacitor paper ed. Indeed, at present a life test appears to be the only practical eans for evaluating capacitor paper, since, within the limits observed in commercial material, the chemical and physical tests ually made do not correlate with life. Lack of correlation is cribed to obscure physical factors which have not yet been ideied.

Generally, several weeks are required to evaluate a paper by life tests of the usual severity. Unfortunately, the duration of these tests is too long for quality control of paper.

The desire for a life test which requires no more than a day or two for evaluation led to the development of a rapid dc test. The philosophy of rapid life testing is based upon the experimental evidence that the process of deterioration under selected temperature and voltage conditions is principally of a chemical nature, and also upon the well-known fact that rates of chemical reaction increase exponentially with temperature.

Life tests on two-layer capacitors conducted at 130°C provide an acceleration in deterioration many fold more than that obtained in the lower-temperature life tests, and correlate well with these tests.

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Introduction

PAPER capacitors, like other components of electronic apparatus, must continue to face demands for higher-temperature and higher-voltage operation, for more dependable service, and for greater compactness. Enormous improvements in dependability and general quality level of paper capacitors have been made over the past several years. Nevertheless, where dependability and long service life are paramount considerations, designs must still be based on the minimum expected life, a value which may be only a few per cent of the maximum expected life.

Our work has indicated that large variation in the durability of paper capacitors subjected to direct voltage results from uncontrolled variables in the paper. Many chemical and physical tests have been proposed for determining the durability of capacitor paper. Our experience with such tests indicates that they may suffice to eliminate the extremely poor papers but are valueless for choosing between, e.g., a moderately poor paper and an excellent paper. In Table I, it is seen that papers of widely different accelerated life test performance exhibit chemical and physical properties which are either identical or which fail to correlate with the life. In our opinion, the reason for this is that the life characteristics of paper are very sensitive to variations in its microstructure which none of the usual chemical and physical tests detects. Pending discovery of tests which will detect such variations, it is necessary to use life tests to determine the voltage performance of paper.

The goal which was set was that of an accelerated life test which could be made in an elapsed time similar to the chemical and physical tests, i.e., a day or two. The result has been a testing procedure requiring about a day for preparation of samples and about the same time for performing the life test.

It is generally agreed that increasing the voltage to accelerate life testing gives unreliable results at voltages much above two and one-half times the rated voltage. Owing to the chemical nature of the deterioration process, high acceleration may be gained through increasing the test temperature.[1]

Equipment

The method developed consists of subjecting two-layer capacitor test units to a direct voltage life test carried out at 130°C and 560 volts per mil. The equipment consists principally of a variable voltage rectifier producing potentials up to 2,000 volts, an oven, and a measuring and control circuit all housed in a steel cabinet. Fig. 1 is a photograph of the apparatus displaying the general arrangement of the assembly and locations of the important components. Each capac on test (Fig. 2) is in circuit with one pole of a dou pole breaker. Facilities are provided for testing capacitors at a time. No units are substituted for fa

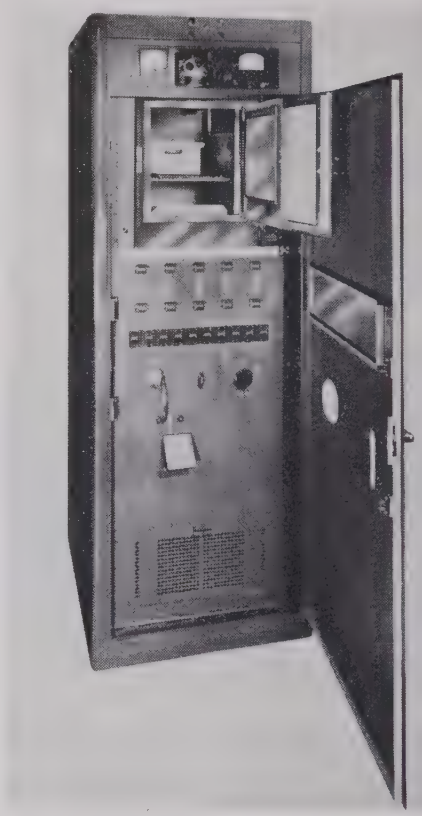


Fig. 1—Direct-voltage performance test set.

specimens. Failure is marked by a steep rise in the c rent in the capacitor circuit. When this current reac about 0.025 ampere the trip circuit is energized, rem

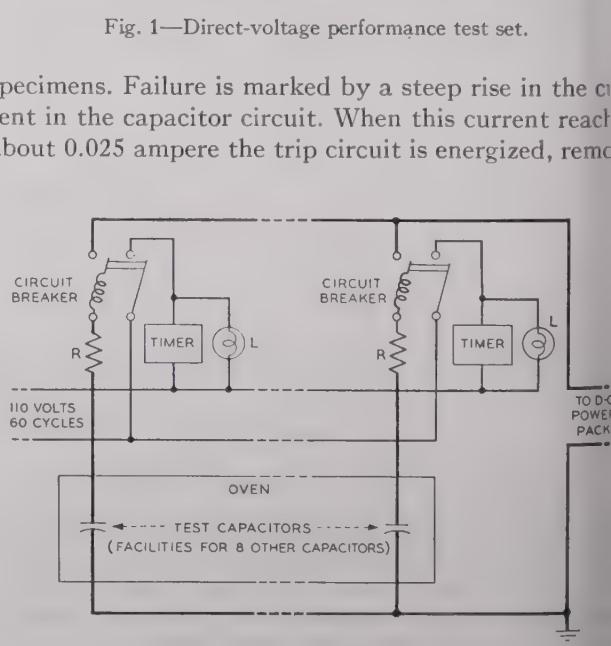


Fig. 2—Test circuit schematic.

ing the voltage from the failed specimen and opening t 110V-60 cps timing circuit associated with the oth

[1] D. A. McLean, L. Egerton, G. T. Kohman, and M. Brotherton, "Paper dielectrics containing chlorinated impregnants. Deterioration in dc fields," *Ind. and Eng. Chem.*, vol. 34, pp. 101; January, 1942.

le of the breaker. The timer is a motor-driven counter ich records the time to one-tenth of an hour. The np circuit in parallel with the timing circuit aids in termining the status of the test.

Description of the Test

If a life test on capacitors is to be used to evaluate pacitor paper, all other variables must be controlled. is means faithful reproduction of conditions of wind-, drying, and impregnation, together with constancy d purity of the other constituents; namely, the im-egnant and the electrode metal.

In this investigation, capacitor units are wound with o layers[2] of paper $2\frac{3}{4}$ inches wide and with 0.00025 h thick aluminum foil electrodes to provide a capaci-ice, when impregnated, of approximately 1.0 μf for 004 inch thick paper. For thicknesses less than 0.0004 h, the windings are constructed to provide a test on same area of paper as contained in a 1-μf 0.0004 h unit. For thicker papers, in order to avoid excessive ysical size, less area is employed.

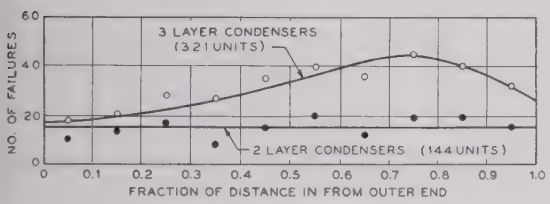


Fig. 3—Distribution of points of failure within test capacitor.

he choice of units of these dimensions resulted from ensive auxiliary investigation in connection with the essity of determining whether or not overheating de-ps within the unit during testing under these condi-is, causing premature failure. Statistically, the weak-point, and hence the point of failure, in the paper et should occur with equal probability on any sur-element of the sheet. Post-mortem examination of rge number of specimens has shown that the two-er units of the above specification exhibit a random ribution of failure along the length of the sheet, in-ating the absence of local internal heating. On the er hand, when tested at the much higher voltages uired to produce failure in a short time, local over-ting occurred in three-layer units of the same physi-limensions as evidenced by an observed peak in the ribution curve, Fig. 3. Another factor which has ored the choice of two-layer units is that such a test been found to be more discriminating than one on e-layer units, producing a greater difference in life n sample to sample. Another major reason is that the bulk of capacitors for telephone use is of two-layer construction.

In the processing and testing, apparatus which is clean and is inert toward the impregnating compound is essential. Materials such as stainless steel, aluminum, nickel- or tin-plated steel, and glass are preferred for test jigs and containers.

The units in Fig. 4 are held under spring pressure between aluminum or stainless plates during all processing and testing. This provides a mechanical pressure of about 12 pounds per square inch. The processing is conveniently considered in three steps: air-baking, vacuum-drying, and impregnating.

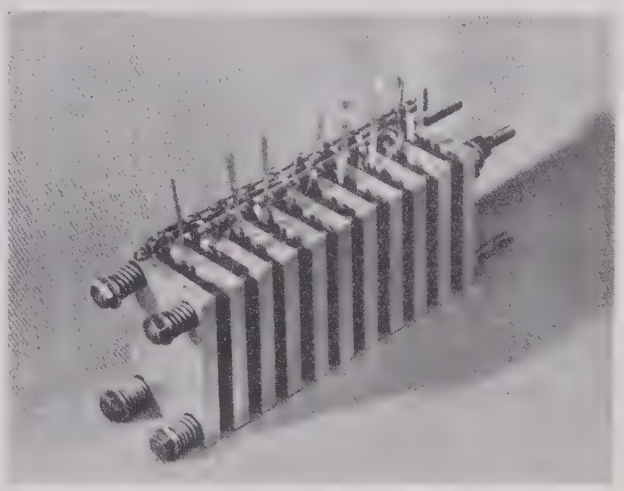


Fig. 4—Capacitor units assembled in clamping jig.

The purpose of the air-bake is to produce limited oxidation of the paper to improve its electrical properties.[3] Also, during the air-bake, the bulk of the moisture is removed. The air-bake is carried out at atmospheric pressure and 130°C for 16 to 18 hours. This temperature is also maintained throughout the following steps.

In the vacuum drying, which follows the air-baking immediately, the units are held under vacuum (<1 mm Hg) for 4.5 hours. During the first 3.5 hours, the vacuum is released momentarily every 30 minutes by admitting dry air. This aids in removing residual moisture. During the last hour the vacuum is applied continuously. The resulting residual moisture content is about 0.05 per cent or less based on the dry weight of the paper tested.

The units are immersed in hot impregnant (130°C) for two hours directly following the drying and without breaking the vacuum. During the first hour, vacuum is maintained. The last hour is at atmospheric pressure.

[2] Two sheets of paper, one superimposed upon the other, are sand-ed between aluminum foils.

[3] D. A. McLean, "Paper capacitors containing chlorinated impregnants. Benefits of controlled oxidation of the paper," *Ind. and Eng. Chem.*, vol. 39, p. 1457; November, 1947.

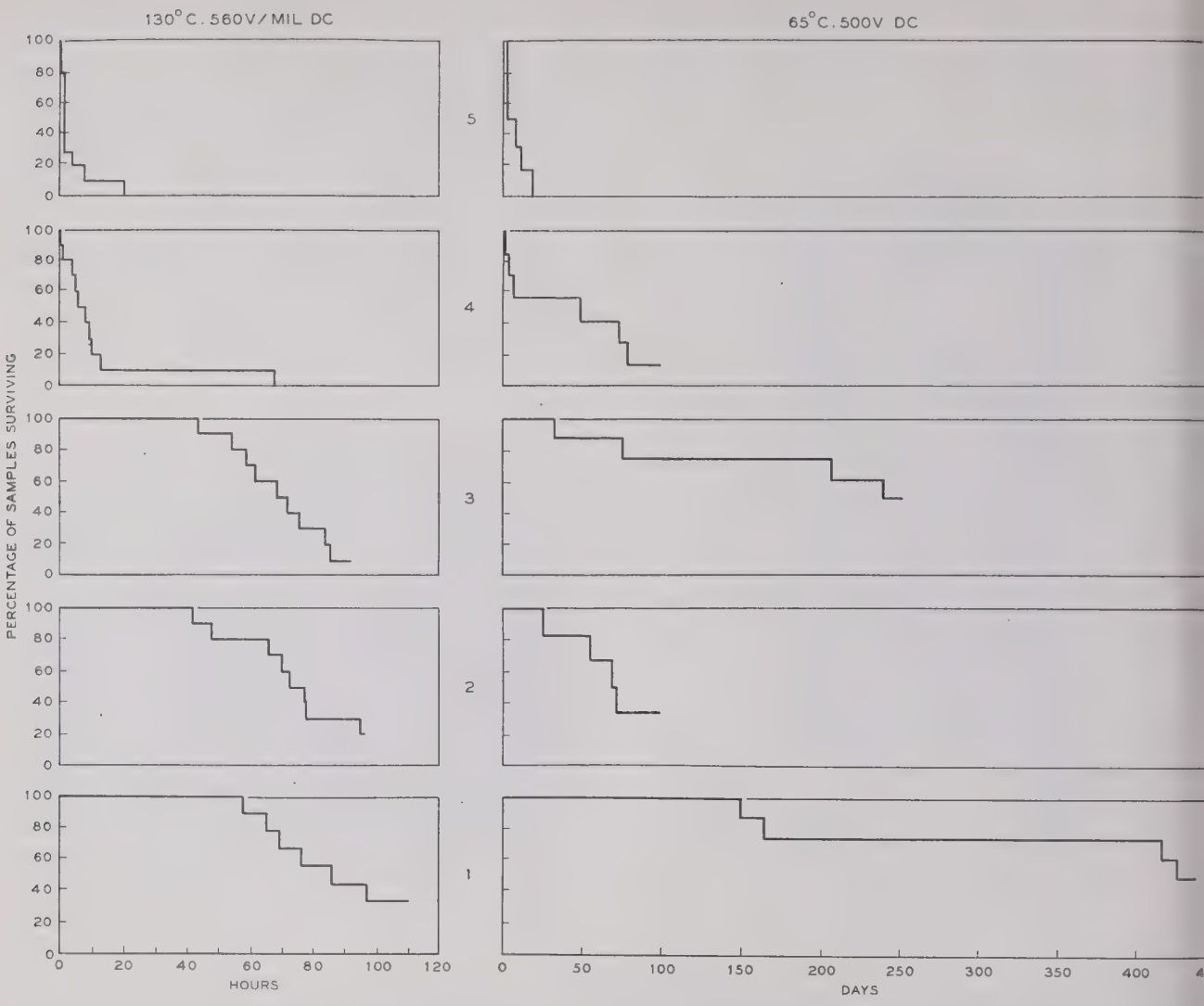


Fig. 5—Comparison of two-layer, 0.4-mil kraft capacitor papers listed in Table I.

The impregnant used in these tests consists of chlorinated naphthalene (Halowax 1001) stabilized with 0.5 per cent of anthraquinone.[4]

A set of test conditions could probably be found which would involve neither stabilization nor air-bake. However, the process described here closely simulates that employed in the manufacture of telephone capacitors.

The test jig with impregnated units in place is then rapidly transferred to the test container and covered with the same compound used in impregnation. Electrical connections are made to the container terminals. Solder connections have been avoided in order to eliminate the possibility of solder flux contamination of the impregnant. The cover is attached and the contain placed in the oven. After equilibration at test tempe ture, the capacitance of each unit is measured as a c cuit check, followed by a short time direct-voltage electric strength test performed by applying a volta about 50 per cent in excess of the life test voltage eliminating mechanically defective samples. The dire current leakage at 100V is then recorded before the vo age is raised to the life test potential.

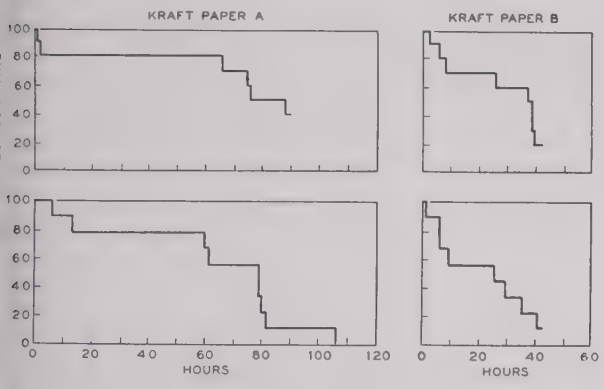
Discussion of Life Test Results

The 130°C life test results on papers referred to Table I are compared in Fig. 5 with tests conducted 65°C and 500 volts. In the step-wise plot the time a in the 130°C life test is hours; in the slower life tests is days. The scales of the time axes are in the ratio 48:1 and it will be observed that the 130°C life test p

[4] D. A. McLean and L. Egerton, "Paper capacitors containing chlorinated impregnants. Stabilization by anthraguinone," *Ind. and Eng. Chem.*, vol. 37, p. 73; January, 1945.

es additional acceleration in deterioration of the or- of the time scale ratio.

The time to fifty per cent failure for both the 130°C l 65°C tests has been recorded in Table I as an arbi- ry yardstick of comparison. The five papers show the ne relative order on both tests with the exception of ersal of the second and third place. The order (1 to 5) shown is on the basis of the results of the 130°C tests. Satisfactory duplicability on check runs, shown on . 6, is readily obtained. The important aspects of plicability are the assurance that the equipment is forming satisfactorily, and that test samples are be- prepared in a consistent manner. Early failures arise n various causes; partly, from mechanical defects in the wound unit, but principally from variability of the paper. Apart from this, it will be observed that the course of the life chart results in comparable total integrated life for duplicate runs, the criterion upon which duplicability is based.



6—Duplicability of results by the 130°C life test 560 volts/mil dc on same paper and same quantity of units

The spread of results that can be expected of commercial capacitor paper can be judged approximately from Fig. 5. The setting up of criteria for acceptable paper is beyond the scope of this paper. However, it appears that in a period of twenty-four to forty-eight hours the general quality level of the lot being tested can be judged, provided that the limited number of test units which it is practical to test is representative of the lot. Sampling, required by statistical considerations, to accomplish this end to a high degree has not been determined.

It appears that this test will be of value in research on capacitors, both in improving the quality of the paper and in isolating the factors which govern the life of capacitors.

For purposes of illustration, data have been given on only 0.0004 inch papers. Extensive investigation indicates that similar relationships hold for other thicknesses of paper.

Acknowledgment

The authors are indebted to J. R. Weeks of these Laboratories for numerous helpful suggestions, and for independent corroborative evidence using equipment of the same design.

TABLE I

Chemical and Physical Properties of 0.0004-Inch Kraft Capacitor Papers

Paper Sample	M-86	M-77	M-89	M-95	M-87
	(1)*	(2)*	(3)*	(4)*	(5)*
Chemical Properties:					
Conductivity of water extract (micromhos/cm)	4.6	7.5	5.0	2.7	5.5
Chloride Content (%)	0.002	0.002	0.003	0.002	0.003
Ash (%)	0.28	0.37	0.36	0.28	0.33
Reaction of water extract (meq/gm)	0.002 Alk	0.003 Alk	0.0001 Alk	0.0002 Acid	0.011 Alk
Water soluble (%)	0.23	0.30	0.44	0.39	0.30
Alcohol soluble (%)	0.06	0.18	0.14	0.16	0.10
Benzene soluble (%)	0.06	0.10	0.04	0.05	0.04
pH	7.5	8.4			
Permanganate number	11.5	12.7	12.7	13.3	12.3
Pentosans	6.9	7.2	7.9	8.0	7.9
Physical Properties:					
Porosity (Average)	1.6	1.8	0.4	0.4	2.6
Porosity (Max. in 15 values)	3.7	3.9	0.8	0.6	7.9
Apparent Density (Gm/cc)	0.94	0.94	0.98	0.99	0.94
Conducting Particles (per sq. ft.)	2.4	2.6	0.6	5.9	3.4
Life (2-layer units—Halowax plus ½% anthraquinone):					
Days to 50% Failure (65°C.–625V/mil dc)	427	72	240	50	8
Hours to 50% Failure (130°C.–560V/mil dc)	86	73	69	8	1

Order as shown on Fig. 5.

Some Aspects of Cathode-Follower Design at Radio Frequencies*

FRED D. CLAPP†, MEMBER, IRE

Summary—Pertinent high-frequency design data including the circuit gain, the gain phase angle, the input impedance in resistive and reactive components, and the maximum allowable input signal voltage may be determined quickly from simple design charts derived by approximations which are applicable over a wide range of frequency and circuit parameters. The only circuit quantities required for use of the charts are the magnitude and phase angle of the cathode load impedance, the internal tube capacitances, and the grid-to-plate transconductance of the tube at the operating point.

A discussion of the effects of the grid-to-cathode capacitance on circuit operation at high frequencies is presented, including a critical analysis of the source impedance. It is shown that the source impedance may be greatly affected by the impedance of the grid driving source unless remedial measures are taken. Various circuit changes which reduce or eliminate the undesirable effects of the grid-to-cathode capacitance are offered with discussion and analyses.

I. Gain Function for a Complex Load Impedance

REFERRING TO Fig. 1, if the load impedance is a resistance R_k, the gain of the cathode follower is the ratio of e_0 to E_g and has been given by many investigators[1,2] as

$$A = \frac{\mu}{1+\mu} \cdot \frac{R_k}{R_k + \dfrac{R_p}{1+\mu}} \tag{1}$$

where

A = gain of stage
μ = amplification factor of tube
R_p = plate resistance of tube
R_k = cathode load resistance.

Equation (1) is easily derived by application of the equivalent plate-circuit theorem.

It is apparent in (1) that if μ is large compared to unity considerable simplification is possible, resulting i

$$A = \frac{1}{1 + \dfrac{1}{R_k G_m}} \cdot \tag{2}$$

If the cathode load impedance is complex, R_k in equation (2) becomes $Z_k = R_k \pm jX_k$. Rationalizing and simplifying the resulting expression, the magnitude of the gain A and its phase angle α (which is the phase angle between e_0 and E_g in Fig. 1) can be shown to be given by the following expressions

$$|A| = \frac{G_m |Z_k|}{\sqrt{1 + 2G_m |Z_k| \cos\theta + (G_m |Z_k|)^2}} = \frac{\sin\theta}{\sin\alpha} \cdot \frac{1}{\dfrac{1}{G_m |Z_k|} + G_m |Z_k| + 2\cos\theta} \tag{3}$$

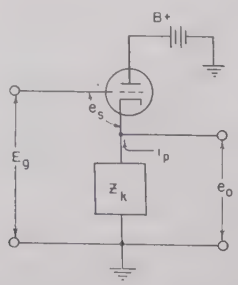


Fig. 1—The gain of the cathode follower is the ratio of e_0 to E_g when R_k is the load impedance.

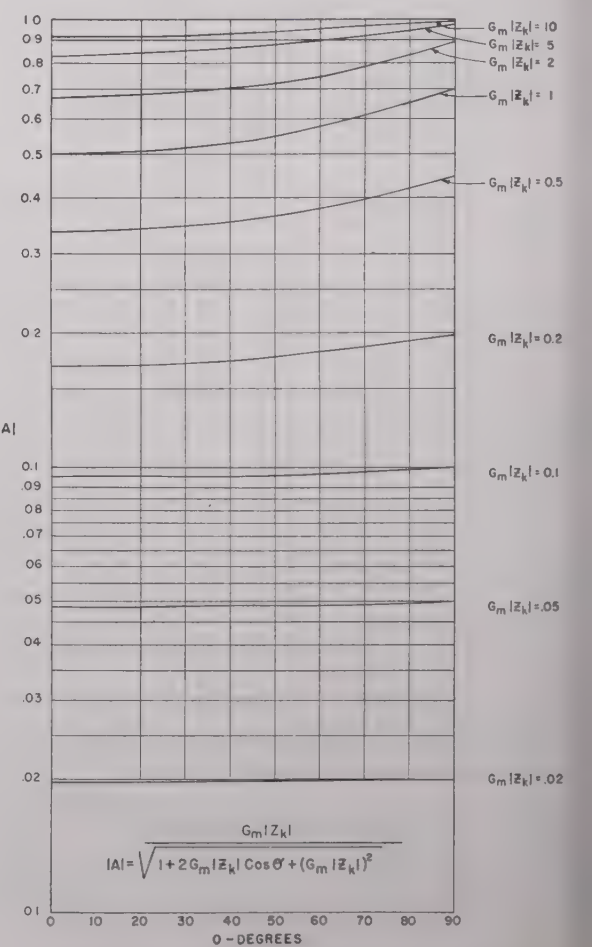


Fig. 2—Magnitude of complex gain A/α as a function of cathode load phase angle θ.

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† Antenna Laboratory, University of California, Berkeley, Calif.

[1] K. Schlesinger, "Cathode follower circuits," PROC. I.R.E., vol. 33, pp. 843–855; December, 1945.

[2] J. Diamond, "Circle diagrams for cathode followers," PROC. I.R.E., vol. 36, pp. 416-420; March, 1948.

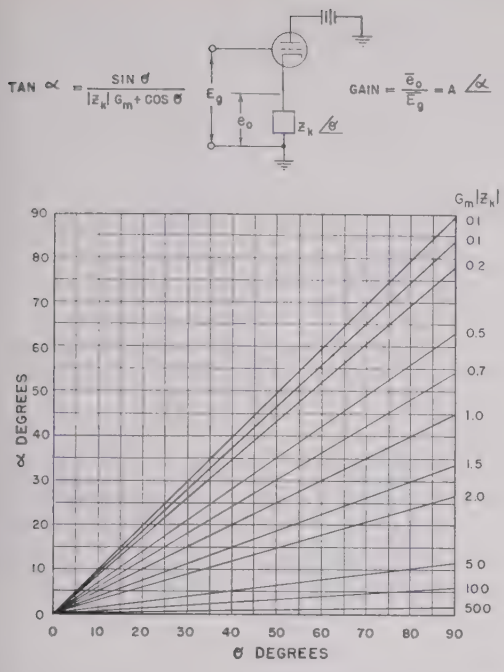

Fig. 3—Relation of cathode load phase angle to angle of omplex gain function for various values of parameter $G_m|Z_k|$.

ere

g_m = grid-to-plate transconductance

θ = phase angle of cathode load impedance

θ = arc tan X_k/R_k

$A|$ = magnitude of complex gain.

he phase angle of the complex gain is

$$\alpha = \text{arc tan} \frac{\sin \theta}{G_m|Z_k| + \cos \theta}. \tag{4}$$

quations (3) and (4) are plotted as Figs. 2 and 3 with product $g_m|Z_k|$ as a parameter. These charts permit quick and easy evaluation of the magnitude and se angle of the stage gain for any load impedance. ow frequencies where the load impedance is essen- y resistive, it may be considered that $\theta = 0$ and the ordinate of Fig. 2 will give the results obtained from iously published nomographs[3] for these conditions. then zero, of course.

II. Effects of Grid-to-Cathode Capacitance

t high frequencies, the grid-to-cathode capacitance ommonly used tubes acts as a coupling impedance of us proportions between input and output circuits. eral first-order results of this coupling are observed. t, the input impedance (dynamic impedance be- n grid and ground) which is normally equivalent ne grid-to-plate capacitance shunted by a portion he grid-to-cathode capacitance and an extremely resistance, is lowered significantly. Depending the phase angle of the cathode load impedance, esistive component may become negative in sign, ough the reactive component is always capaci-

[3] M. B. Kline, "Cathode follower nomograph for pentodes," *Elec- s*, vol. 20, p. 136; June, 1947.

tive. With capacitive cathode loads of suitable proportions, the negative resistance seen looking into the grid circuit may become so low that oscillation in some mode is virtually a certainty. The input capacitance may be significantly increased. Using the symbols of Fig. 1, the input current may be written:

$$i_g = \frac{\bar{E}_g - \bar{e}_0}{-jX_{gk}} \tag{5}$$

$$i_p \gg i_g$$

where i_g = current flowing to the grid. By previous definition,

$$e_0 = E_g A \underline{/\alpha}. \tag{6}$$

Then

$$Z_g = \frac{\bar{E}_g}{\bar{i}_g} = \frac{-jX_{gk}}{1 - A\underline{/\alpha}}. \tag{7}$$

In deriving (7), nothing has been said about the current flowing into the grid-to-plate capacitance. This component of the impedance is a static quantity unaffected by the tube operation (since the plate is assumed to be grounded for rf) and may be added in later.

For convenience, (7) may be rationalized to separate the resistive and reactive components. Performing this operation yields

$$Z_g = |X_{gk}| \left[\frac{\mp |A| \sin \alpha - j(1 - |A| \cos \alpha)}{1 - 2|A| \cos \alpha + |A|^2} \right]. \tag{8}$$

The $\mp$ sign corresponds to that of the cathode-load phase angle θ.

Equation (8) expresses the new components of input impedance due to coupling of the grid-to-cathode capacitance as a series circuit of capacitance and resistance across the input terminals. It is more convenient to express this new impedance in terms of an equivalent parallel circuit of resistance and capacitance. This is easily done with the standard circuit transformations, yielding

$$R_{par} = \frac{X_{gk}}{|A| \sin \alpha} = \frac{X_{gk}\sqrt{1 + 2G_m|Z_k| \cos \theta + (G_m|Z_k|)^2}}{A \sin \theta}. \tag{9}$$

$$X_{par} = \frac{X_{gk}}{1 - |A| \cos \alpha} = \frac{X_{gk}}{1 - |A| \dfrac{G_m|Z_k| + \cos \theta}{\sqrt{1 + 2G_m|Z_k| \cos \theta + (G_m|Z_k|)^2}}}. \tag{10}$$

$$C_{par} = C_{gk}(1 - |A| \cos \alpha) = C_{gk}\left[1 - |A| \frac{G_m|Z_k| + \cos \theta}{\sqrt{1 + 2G_m|Z_k| \cos \theta + (G_m|Z_k|)^2}}\right] \tag{11}$$

where

R_{par} = equivalent shunt resistance appearing across input terminals

X_{par} = equivalent shunt reactance appearing across input terminals

C_{par} = equivalent shunt capacitance appearing across input terminals.

Equations (9), (10), and (11) are normalized with respect to X_{gk} and C_{gk} and plotted in Figs. 4, 5, and 6. After first finding $|A|$ and α from the curves previously presented, the input impedance components may be very quickly found with sufficiently good accuracy for most engineering purposes.

Equation (9) shows that for frequencies and tube capacitances such that X_{gk} is a few hundred ohms, the cathode follower will make a vigorous oscillator if the cathode load is capacitive so that R_{par} has a negative sign. This phase of operation has been well covered in several of the references.

The second very important result of the grid-cathode coupling is its influence upon the effective source impedance of the cathode follower.

The source impedance may be defined as that impedance seen looking into the output terminals toward the tube as in Fig. 7. Here a shunt circuit of three elements is seen represented by the cathode load impedance Z_k, the tube plate resistance R_p in series with the equivalent

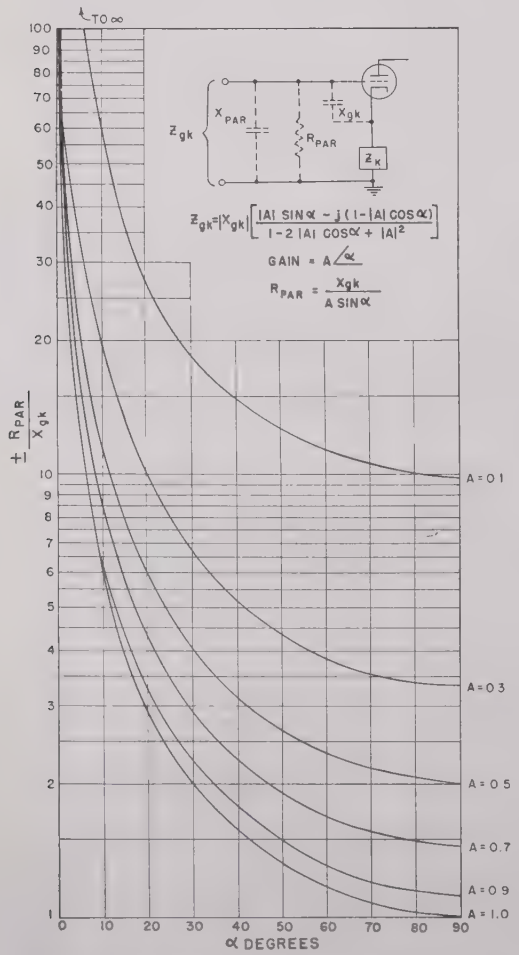

Fig. 4—Equivalent shunt resistance of cathode-follower input impedance.

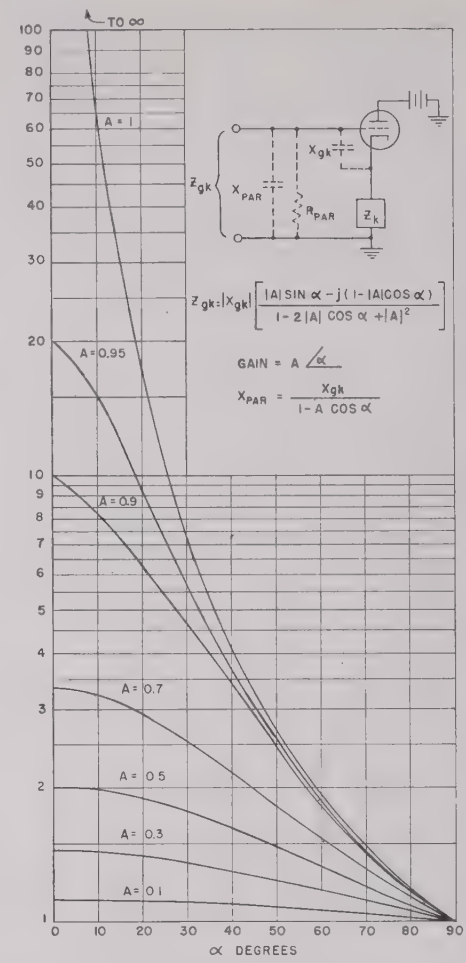

Fig. 5—Equivalent shunt reactance of cathode-follower input impedance.

generator, and the cathode-grid reactance in series wi whatever impedance appears between grid and grou represented by the symbol Z_{gg}. It is important to no that Z_{gg} may be anything from a very low value to very high value depending on the nature of the drivi source and stray reactances.

The circuit of Fig. 7 may be solved by Kirchhof laws from the equations

$$i_t Z_k - i_p Z_k - i_g Z_k = E_t, \quad (1$$

$$i_p R_p = E_t + \mu e_s, \quad (1$$

$$i_g(Z_{gg} - jX_{gk}) = E_t, \quad (1$$

where E_t = test voltage applied across output termina

For convenience, a new quantity β will be defin which is given by

$$\beta = \frac{-jX_{gk}}{Z_{gg} - jX_{gk}}. \quad (1$$

β represents the portion of cathode-ground volta E_t which appears between grid and cathode e_s due to sta ic voltage division by the impedances from grid cathode and from grid to ground.

Simultaneous solution of (12), (13), and (14) and i sertion of (15) gives the source impedance as

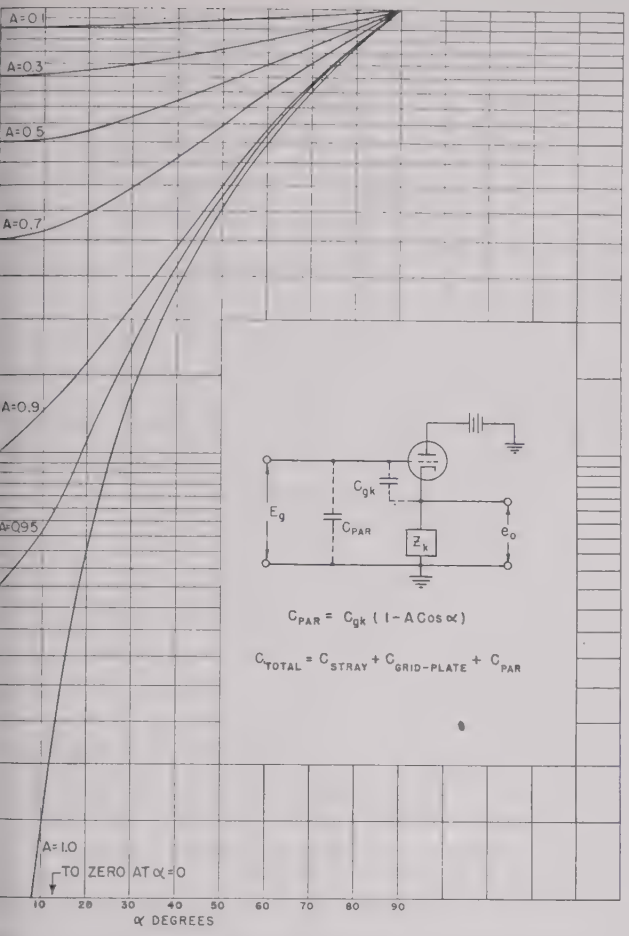

Fig. 6—Ratio of apparent input capacitance to grid cathode capacitance as a function of cathode follower complex gain A/α.

$$Z_s = \frac{1}{\frac{1}{Z_k} + \frac{1}{Z_{gg} - jX_{gk}} + \frac{1+\mu\beta}{R_p}} \cdot \qquad (16)$$

approximations are involved in this equation except assumption that all the voltage across Z_k is due to plate current; i.e., $i_p \gg i_g$.

nspection of the circuit reveals that the first 2 de- ninator terms are static quantities unaffected by tube ration. The term $(1+\mu\beta)/R_p$ corresponds to the dy- nic component of the source impedance which is of ticular interest. If $\beta=1$, this term is seen to have a ue so high with most tubes that the other denomina- terms may be neglected and the source impedance tten in the form usually given for low frequencies;

$$Z_g\Big|_{\text{low frequency}} = \frac{R_p}{1+\mu} \cong \frac{1}{G_m} \cdot \qquad (17)$$

Equation (15) shows that, if the grid-to-ground im- ance is comparable to the grid-to-cathode imped- e, (17) is no longer a good approximation. This is y likely to happen in the case where a tuned circuit onnected between grid and ground so that the stray ctances are tuned out, becoming part of a tuned cir- at parallel resonance. In this case, however, the l-to-ground impedance may assume values so large

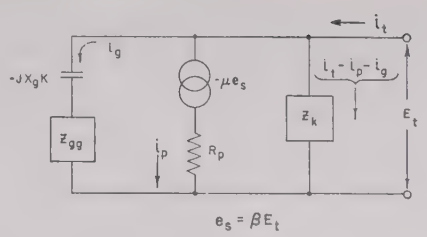

Fig. 7—Shunt circuit represented by the cathode load impedance Z_k and the cathode-to-grid reactance in series with impedance Z_{gg}.

compared to X_{gk} that β approaches zero and the term $(1+\mu\beta)/R_p$ approaches $1/R_p$. The source impedance then becomes quite large and the expected advantages of the cathode follower fail to materialize.

From the foregoing, it is clear that, if a cathode follower is driven by a tuned grid circuit at frequencies of the order of megacycles, the operation is likely to be unsatisfactory. The output will be lower than expected due to the increased source impedance and oscillation is likely at either the tuned circuit frequency, the input leads frequency, or both. This presupposes that the cathode load impedance will be capacitive, a probable result of the cathode-to-heater capacitance and other strays.

III. Remedial Measures

If operation with a tuned grid circuit is required, several methods are available for nullifying, or at least reducing, the effects of grid-to-cathode coupling.

(a) The oscillation may be suppressed by inserting a resistor of suitable size in series with the grid to damp the circuit. Depending on the frequency, this resistor will load the tuned circuit in a manner to be analyzed later. This method has definite limitations.

(b) A coil may be placed from grid to cathode, with a suitable dc blocking capacitor to resonate the grid-to-cathode capacitance (Fig. 8). This increases the value of X_{gk} by a very large amount and is effective in lowering the source impedance and suppressing oscillation at the design frequency. However, oscillation at frequencies determined by the grid-circuit leads may still occur since Z_{gk} is high at only one frequency. This method of operation is subject to the serious disadvantage that the coil must be changed for each change of operating frequency.

It is, of course, possible to make the blocking capacitor C_B variable so that the total reactance may be ad-

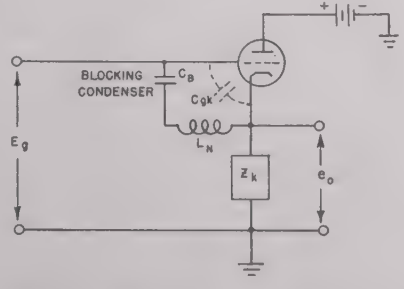

Fig. 8—Blocking capacitor C_B in series with L_N to resonate the grid-cathode capacitance C_{gk}.

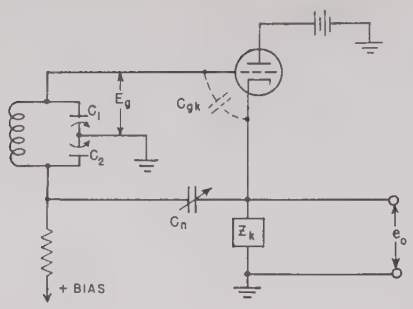

Fig. 9—Bridge neutralization using grid-to-cathode capacitance as one arm.

justed to the correct value at each operating frequency.

(c) The analogy of the foregoing to the familiar "coil neutralization" used in fixed-frequency broadcast and other transmitters leads to the circuit of Fig. 9 using the grid-to-cathode capacitance as one arm of a bridge circuit. Variations of this circuit are those of Figs. 10 and 11.

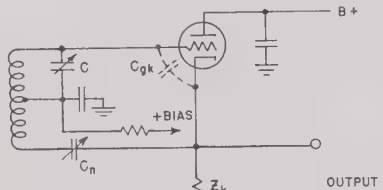

Fig. 10—Alternate method of coil neutralization from that used in Fig. 9.

To analyze the effects of these circuits upon the source impedance, refer to (16). It is seen that the only variable is β and that if it can be kept close to unity the source impedance will remain low. Remembering that β was defined as the ratio of grid-cathode to grid-ground voltages, we may draw the equivalent circuit of the cathode follower with bridge neutralization, look into the output terminals toward the tube, and deduce the source impedance. The equivalent circuit is given in Fig. 12. It is unnecessary to consider Z_k in the equations since it is across the bridge and does not disturb the ratio of the arms. Referring to Fig. 13, the equations to be solved are

$$Z_{gk}i_2 + Z_1 i_3 = e_{in} \tag{18}$$

$$Z_n i_1 - (Z_n + Z_L + Z_{gk})i_2 + Z_L i_3 = 0 \tag{19}$$

$$Z_2 i_1 + Z_L i_2 - (Z_2 + Z_L + Z_1)i_3 = 0. \tag{20}$$

Simultaneous solution of the above yields the currents

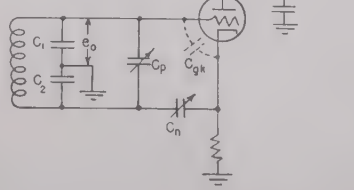

Fig. 11—Another method of coil neutralization than that used in Figs. 9 and 10.

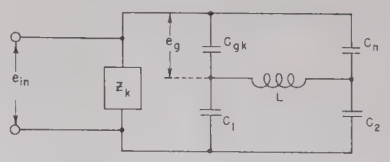

Fig. 12—Equivalent circuit of the cathode follower with bridge neutralization.

which may then be used to find the desired ratio of voltages. This is

$$\beta = \frac{e_g}{e_{in}} = \frac{C_1 + C_2 + j\omega Z_L C_1 (C_n + C_2)}{C_1 + C_2 + C_{gk} + C_n + j\omega Z_L (C_1 + C_{gk})(C_n + C_2)}. \tag{21}$$

The conditions for balance of the bridge are that the ratio C_{gk} to C_n should be the same as the ratio of impedances on each side of the tank-circuit ground tap. This is expressed by

$$\frac{C_1}{C_2} = \frac{C_{gk}}{C_n}. \tag{22}$$

If (22) is substituted in (21) there results

$$\beta = \frac{C_1}{C_1 + C_{gk}}. \tag{23}$$

This simple relation is seen to contain no frequency term. Since the tuning capacitor C_1 is nearly always much larger than the grid-to-cathode capacitance C_{gk}, β is nearly always close to unity.

If the circuit of Fig. 11 is used, the output voltage may be increased by making $C_2 = nC_1$ and $C_N = nC_{gk}$ where $n > 1$. Equation (23) still holds.

Theoretically, for perfect balance to exist it is necessary that the stray reactances on both sides of the tank circuit be balanced, and that if a resistor is used in series with the grid another be placed in series with the neutralizing capacitor. In practice, sufficiently good results are usually obtained without these precautions. With any of the three neutralizing circuits shown (Figs. 9, 10 or 11) measurements indicate that the source impedance is constant to about 10 per cent when the grid circuit is tuned through resonance. Even better operation can be obtained at one frequency by careful adjustment.

Circuits in which the *coil* center tap is grounded for rf should be avoided as they will tend to oscillate at the resonant frequency of the circuit between grid and ground, usually 20 or 30 Mc. Parasitics in the split capacitor circuits shown (Figs. 9 and 11) usually occur at frequencies determined by the lead path from the grid through the upper half of the split capacitor to ground.

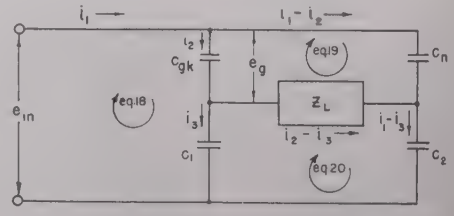

Fig. 13—Circuit analysis of the currents which are used to find the desired ratio of voltages.

path is normally much shorter so that the oscilla- has a frequency of several hundred Mc and is much vigorous. In all split-capacitor circuits tested in the uency range from 2 to 30 Mc. a grid resistor of 47 s was sufficient for complete stability, and a value) ohms was sufficient in many cases.

is important to keep this resistor as small as possi- since it may have a serious effect on the input cir- Q at the higher frequencies. This is due to the fact the grid circuit voltage causes a current to flow ugh a path composed of the grid-cathode capaci- e, the neutralizing capacitance, and the grid re- r. The loss due to this current results in a lowering ie Q of the input circuit. The amount of loss is di- y proportional to frequency squared since the cur- is proportional to frequency.

y simple transformations, it is possible to postulate quivalent resistance in series with the tank circuit lucing the same loss. The working Q of the circuit then be expressed in terms of the unloaded Q, the uit parameters, and the frequency. The relation is

$$= \frac{\omega L}{\frac{\omega L}{Q} + (R_g + R_n)\left(\frac{n^2 C_{gk}^2}{[C(1+n) + nC_{gk}]^2}\right)} \qquad (24)$$

$$(R_g \ll X_{gk})$$

re

ωL = reactance of tank coil
Q = unloaded Q of tank coil
R_g = grid resistor
R_N = resistor in series with neutralizing capacitor if used
C = total effective tuning capacitance
Q_{eff} = effective Q of input circuit
$n = C_n/C_{gk}$.

With commonly used circuit values, a grid resistor of 47 ohms will lower the circuit Q by one half at 20 Mc.

IV. Permissible Input Voltage

One aspect of cathode follower operation not usually stressed in the literature is the effect of a complex cathode load on the magnitude of signal which may be applied to the input circuit.

Consider the case where the cathode load is a complex impedance. Then the input and output voltages have a phase angle α as discussed before, and it is found that the grid-to-cathode voltage, which is the vector difference between the input and output voltages, must of necessity become comparable to those voltages when α assumes moderate values. The magnitude of this effect depends on the value of the gain $|A|$ as well as the angle α, being greatest when $|A|$ is large. It is easily evaluated by setting the the output voltage equal to the input voltage times the complex gain. Thus

$$e_0 = E_g A \underline{/\alpha}. \qquad (25)$$

The grid-cathode voltage e_s is given by

$$\bar{e}_s = \bar{E}_g - \bar{e}_0 = E_g(1 - A\underline{/\alpha}). \qquad (26)$$

Dividing (26) by (25) gives

$$\frac{e_s}{e_0} = \frac{1}{A\underline{/\alpha}} - 1. \qquad (27)$$

Equation (27) is plotted in Fig. 14 for various values of α. Inspection of Fig. 14 shows that for moderate gains of the order of 0.8, the grid-to-cathode voltage becomes equal to the output voltage at $\alpha = 45°$ so that the signal which may be handled under class A conditions is quite limited. The ratio of grid-to-cathode voltage to input (grid-to-ground) voltage is found by dividing the above result by $|A|$.

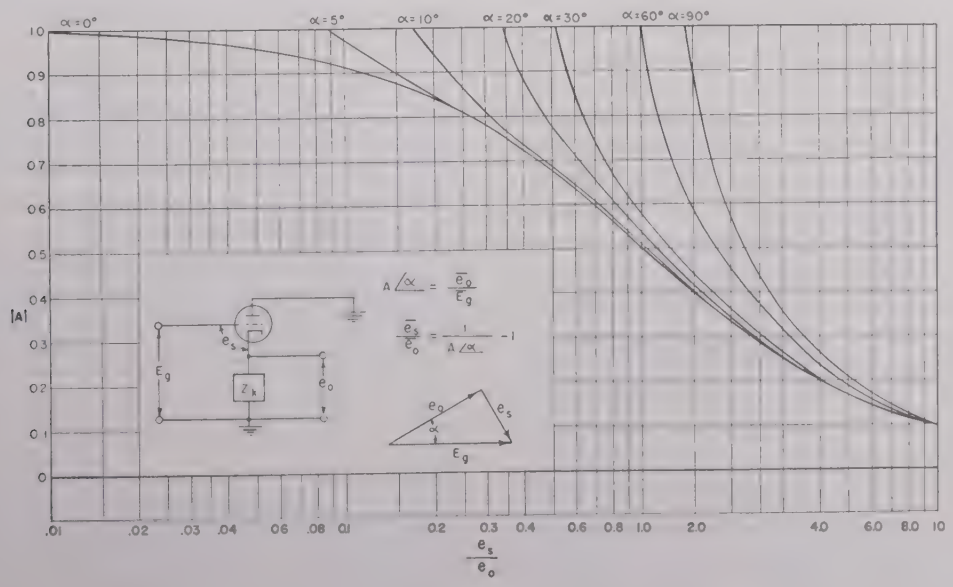

Fig. 14—Ratio of e_s/e_0 for various values of complex gain factor $A\underline{/\alpha}$.

Noise from Current-Carrying Resistors 20 to 500 Kc*

R. H. CAMPBELL, JR.†, STUDENT MEMBER, IRE, AND R. A. CHIPMAN‡, SENIOR MEMBER, IRE

Summary—**Measurements have been made of the effect, first studied by Bernamont that when a direct current is flowing through certain types of resistors, the noise voltage generated at their terminals is considerably in excess of the thermal value. The effect has been investigated for solid carbon-composition, "metallized," palladium film, and "cracked-carbon" resistors, for resistance values from 1,000 to 30,000 Ω, currents from 1 to 10 milliamperes, and at frequencies between 20 and 500 kc.**

For some individual resistors, especially of the solid carbon-composition type, the noise voltage measured at any frequency in the range stated (with direct current flowing) showed much larger fluctuations than those characteristic of thermal noise, with a detecting system having a response up to a few cps. The fluctuations frequently reached peak amplitudes of several times the mean level of the noise. Several records of the fluctuations, taken on a recording meter, are shown. The fluctuations were particularly large when the current was first applied to any resistor, while its resistance changed to a stable value at a higher temperature, suggesting an analogy to the Barkhausen effect in the magnetization of ferromagnetics. No simple correlation of the occurrence of the fluctuations with other factors has been found.

I. Introduction

THAT NOISE in excess of thermal noise is generated by certain types of resistors when a steady current flows through them was first observed in detail by Bernamont in 1934,[1–4] although a brief reference to what was probably the same phenomenon was given by Hull and Williams in 1925,[5] and a similar effect for a mixture of granular carbon particles was mentioned by Kawamoto in 1919[6] and by Frederick in 1931.[7]

Bernamont found the spectral density of this extra noise to be often many times that of thermal noise for the same resistor, and to show a marked decrease with increase of frequency. Since Bernamont, many investigations, both theoretical and experimental, have been made of this type of noise.[8–17] The experimental result have commonly shown that the spectral density of thi noise is proportional to some power of the current clos to 2 (values from 1.5 to 2.7 are reported) and for fre quencies below about 10 kc varies inversely as the firs power of the frequency. At higher frequencies, th spectral density decreases more rapidly.

No name has so far been attached to this additiona component of noise from current-carrying resistors. W will refer to it as current noise throughout this paper.

This paper presents more comprehensive data tha has so far been reported on current noise from severa types of solid carbon-composition and "metallized (i.e., carbon-composition film) resistors in common use showing variation with resistor type, resistance value dc current, and frequency. Measurements have bee made with one-half, 1-, and 2-watt resistors, havin resistance values between 1,000 and 30,000Ω, carryin steady currents of 1 to 10 milliamperes, the noise outpu being measured at frequencies between 20 and 500 kc

There is also described in some detail another nois phenomenon, to which only one slight reference has pre viously been reported,[3] a case of current-noise voltage whose fluctuations about their mean value far exceed the fluctuations characteristic of thermal or shot-effec noise, for the same mean value of noise and otherwis identical conditions of measurement.

II. Apparatus and Method of Measurement

Fig. 1 shows a block diagram of the equipment used i the noise measurements. The resistor under test i mounted (closely shielded) in unit *A*, which contains

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‡ McGill University, Montreal, Que.

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necessary power supply, filters, chokes, etc., to al- passing measured dc currents up to several milli- peres through the resistor, while maintaining a ably high shunt impedance across it at the fre- ncy of measurement. The unit also contains a noise le (British type CV172), with auxiliary circuits, ch is used as a calibration-noise source. Its load is a inductive wire-wound resistance.

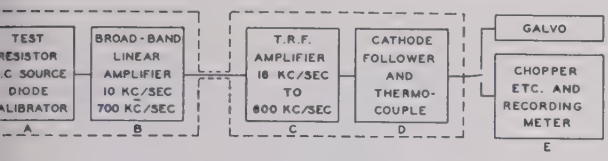


Fig. 1—Schematic diagram of the apparatus used in the noise measurements.

he noise output of unit *A* is amplified by unit *B*, ch is a broad-band linear amplifier with cathode- ower output to unit *C*. The maximum gain of unit s about 33 db over the frequency range 10 to 700 kc. Jnit *C* is a modified low-frequency radio receiver e TE236B, built by RCA Victor Co., Ltd., Canada, the Royal Navy. This tunes over the frequency ge 15 to 600 kc. The 3 tuned amplifier stages result n effective noise bandwidth that varies from about 4 kc over the tuning range of the amplifier, having lower value at the lowest frequency. The calibration hod used in the measurements makes it unnecessary now this bandwidth.

he output of unit *C* goes to a cathode follower in *D*, the output of which is capacitively coupled to heater of a vacuum thermocouple. The output of the mocouple is filtered with a low-pass filter. For mean- are noise-voltage measurements, it is then connected ctly to a galvanometer. To obtain a record of the tuations in the filtered thermocouple output, it is to unit *E*, where it is chopped at 60 cps by a "high- d" relay. The chopped signal is amplified, rectified, red, and fed to a recording milliammeter of maxi- n range 5 milliamperes. While the input to the mocouple is at the frequency to which unit *C* is d, the highest frequency in its output voltage is ted by its thermal time constant, which is of the r of 0.1 seconds. The time constant of the recording er is somewhat higher than this, but less than 0.5 nds. The filters used have low attenuation at fre- ncies below 10 cps. The result is a recording system se response begins to fall off at a frequency of about s and is small at frequencies above 10 cps.

III. Experimental Measurements

Calibration Technique

or any test resistor in unit *A*, if the filtered output age of the thermocouple in unit *D* has only "normal" tuations, this voltage is a quasi-linear function of the n-square noise voltage at the input of the thermo- couple, which noise voltage is made up of contributions from shot effect, flicker effect, thermal noise, and cur- rent noise. However, mean-square noise voltages from various sources are directly additive in producing the resultant mean-square noise voltage, provided the volt- age amplification of the apparatus involved is linear for the range of amplitudes covered.

If, therefore, the test resistor terminals in unit *A* are shorted, the thermocouple output voltage will then be a measure of the noise generated in the amplifier itself, which noise will remain as a constant mean-square com- ponent when additional noise is added. When a test re- sistor is connected, the thermocouple output will in- crease because of the thermal noise contributed by the resistor.

If now a small dc current is caused to flow through the test resistor, it is found in general (but not with wire- wound resistors), that the thermocouple output rises again, the increase in this case being due to a current- noise component. The amount of this current-noise con- tribution can be measured by connecting the diode noise generator in place of the test resistor and adjusting its plate current to give the same two thermocouple outputs that were recorded for the test resistor with and without the dc current flowing. From the well-known calculable noise properties of a saturated diode, the spectral den- sity of the current noise, expressed for example as mean- square volts per 1 cps noise bandwidth, can be found from the values of the diode plate current and the diode load resistor. The gain, frequency, and bandwidth of the intervening amplifiers are not involved in the cal- culation.

This is the method of measurement and calibration that has been used throughout the present work. It means that values given for current-noise spectral den- sity for any resistor are for that component of noise alone.

2. Mean-Square Current-Noise Voltages

The current-noise output of a typical one-half watt "metallized" resistor having a dc resistance of 11,100Ω is shown in Fig. 2 as a function of current and frequency. At any frequency, the variation is as the square of the current, within experimental error. The variation with frequency at any current, however, does not follow a simple law over the frequency range of the measure- ments. At frequencies below 50 kc, the very limited data suggest that the noise spectral density varies as an in- verse power of the frequency somewhat less than unity. At higher frequencies, an inverse power of about 1.6 is indicated.

It is evident that the frequency-variation law is inde- pendent of current, and that the current-variation law is independent of frequency.

Frequency-variation data similar to that of Fig. 2 for 6 different types of resistors of approximately equal dc resistance and all carrying the same value of constant

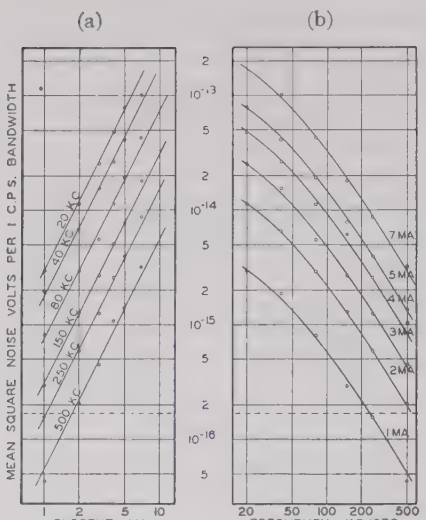


Fig. 2—Spectral density of current noise for a one-half watt "metallized" resistor, dc resistance 11,100 Ω, resistor *A* of Table I. (a) as a function of current at various constant frequencies; the full lines have a slope of 2; the broken line is the theoretical spectral density for thermal noise at room temperature. (b) as a function of frequency at various constant currents; the curves are of identical shape with vertical displacement; the negative slope at low frequencies is slightly less than 1, and at high frequencies is about 1.6.

dc current are given in Fig. 3, which shows that the physical nature of a resistor determines very markedly the magnitude of its current-noise output. The frequency-

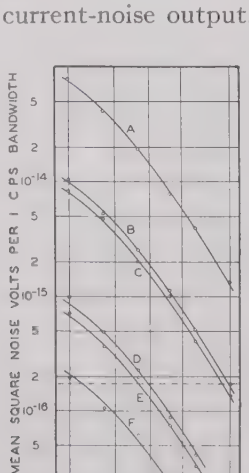


Fig. 3—Spectral density of current noise for 6 different types of resistor (see Table I below) as a function of frequency. The current in each case is constant at 5 milliamperes. The range of dc resistance is from 9,800 to 11,100 Ω.

TABLE I

Resistor	*DC Resistance* Ω	*Wattage Rating*	*Type*
A	11,000	½	"metallized"
B	11,000	1	"metallized"
C	10,300	2	"metallized"
D	9,800	½	carbon composition (Ohmite Little Devil)
E	10,300	1	"cracked carbon"
F	10,200	1	palladium film
G	11,400	½	carbon composition

variation law, however, is apparently identical for different resistors.

Measurements on a large number of each of 3 ty of resistors, and covering a range of resistance valu are plotted in Fig. 4. The frequency and the dc curr

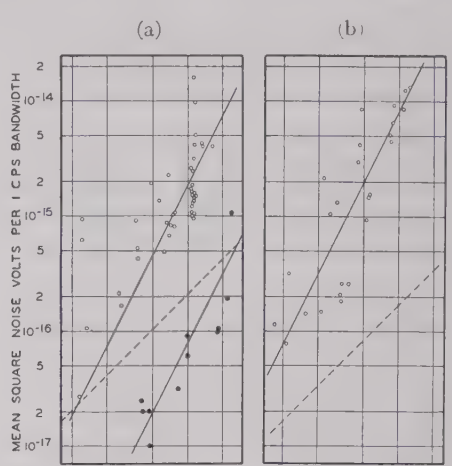


Fig. 4—Spectral density of current noise for 79 resistors of 3 types a function of dc resistance value. The current is 5 milliamp and the frequency of measurement 150 kc in every case. (a) open circles are for 1 watt "metallized" resistors (including sistor *B* of Table I); the black circles are for one-half watt carb composition resistors (including resistor *D* of Table I). (b) half watt "metallized" resistors (including resistor *A* of Table The full lines have a slope of 2. The broken lines show the co sponding theoretical thermal noise level and have a slope of

are constant throughout. This shows that apparen identical resistors (see the points for 20 1-watt "met lized" resistors having dc resistances between 11,000 a 12,000Ω) have current-noise outputs varying by a fac of 20. It is, therefore, not possible to infer with any curacy a resistance-variation law for current noise. T full lines shown correspond to a square law of variati which does appear to fit the data at least better tha line corresponding to a first-power law, which would parallel to the dotted thermal-noise line.

3. *Fluctuations in Current-Noise Voltages*

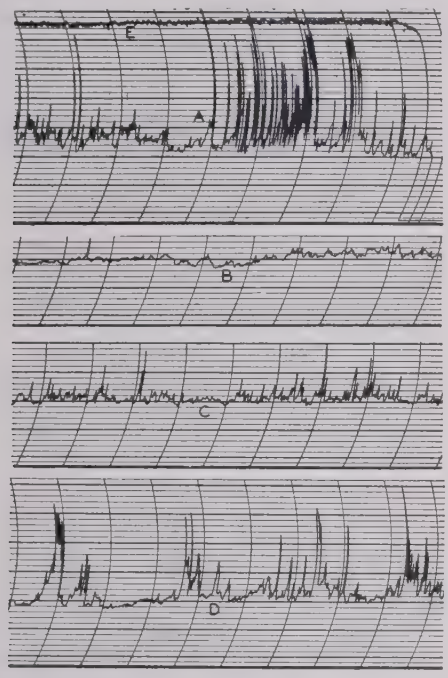
While making the current-noise measurements scribed in the last section, it was found that, for so resistors and currents, the galvanometer indicating t thermocouple output fluctuated wildly. This was marked contrast to its fluctuations when the noise inp to the amplifier was either thermal noise or diode sh effect noise, in which cases its peak fluctuations did exceed about 5 per cent of its mean reading, a figu which is in reasonable quantitative agreement with t theoretical conclusions of Rice[18] who has calculat what the ratio of standard deviation to mean val should be for random noise energy, as a function bandwidth and averaging time for any detector. Wh his calculations are for noise with a spectral density

[18] S. O. Rice, "Filtered thermal noise—fluctuation of energy a function of interval length," *Jour. Acous. Soc. Amer.*, vol. 14, 216–227; April, 1943.

pendent of frequency, the variation of spectral density current noise over the bandwidths used in the present rk is sufficiently small that his conclusions should ply.

Unit *E* of the apparatus, described in section II, was signed to permit comparison of these low-frequency ctuations in the current-noise output of different re-tors, by making graphical records of the thermocouple tput voltage as a function of time.

Fig. 5 shows portions of typical records of these fluc-tions.



5—*A*, *B*, *C*, and *D* are 35-minute sections of a continuous 16-our record of fluctuating current noise from a one-half watt carbon-composition resistor (resistor *G* of Table I). The time scale s from right to left. The current is constant at 5 milliamperes and he frequency of measurement 150 kc throughout. The amplifier ain is the same for all four traces, and for trace *E*.
Trace *A*—the initial 35 minutes of record after connecting the resistor;
Trace *B*—a section showing very low fluctuation;
Trace *C*—a section showing average fluctuation; about one half the 16-hour record was similar to this;
Trace *D*—a section showing large fluctuations; 4 or 5 sections of this type occurred during the 16-hour record.
Trace *E* is for comparison and shows nonfluctuating current oise from resistor *A* of Table I. The vertical scale of the charts, vhile not linear, is indicated roughly by the location of trace *E*, vhich from Fig. 2 corresponds to a noise spectral density of $\times 10^{-15}$ mean-square volts per 1 cps bandwidth.

Trace *E* is a representative section from a continuous our record of the noise output (mainly current noise) n a one-half watt "metallized" resistor of dc resist-e 11,100Ω (resistor *A* of Figs. 2 and 3 and Table I), ying a current of 5 milliamperes. The measurement uency was 150 kc. The fluctuations in this record of the same magnitude as occur in records of shot-ct noise or thermal noise having the same mean level. y may, therefore, be regarded as "normal" fluctua-s.

Traces *A*, *B*, *C*, and *D* are sections from a single con-ious 16-hour record of the noise output from a one-half watt solid carbon-composition resistor, dc resistance 11,400Ω (resistor *G* of Table I). The amplifier gain, the dc current, and the measurement frequency are the same as for trace *E*.

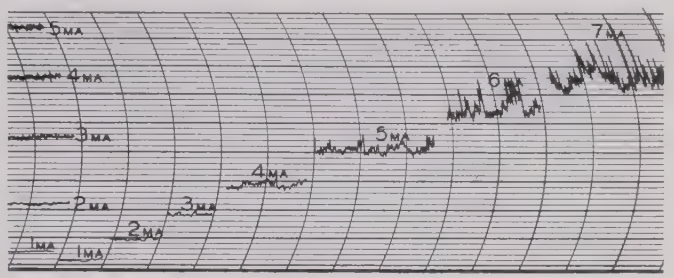


Fig. 6—Fluctuations of current noise as a function of current. The traces at the left are for resistor *A* of Table I, those at the right for resistor *G*. The frequency is 150 kc.

Fig. 6 shows records, each of a few minutes duration, of the noise from the same two resistors for various values of dc current. All the conditions for the 5-milliampere record from resistor *G* are the same (except amplifier gain) as in Fig. 5. The fluctuations in Fig. 6 appear by comparison to be below average for this particular resistor.

A large number of traces taken with these and other resistors and at frequencies from 20 to 500 kc, all other factors remaining constant, show no change in the nature of the fluctuation phenomenon with amplifier frequency for any one resistor. Quantitative data have not been obtained, nor has it been possible to determine whether or not the abnormal fluctuations are simultaneous at different frequencies.

IV. Relation of Experimental Results to Previously Published Work

Attempts to develop theoretical expressions for the spectral density of current noise as a function of current and frequency have most commonly arrived at an equation of the form

$$A(f, I) = \text{constant} \cdot I^2 \cdot \frac{1}{1 + \omega^2\tau^2}$$

where A is the spectral density (for example in mean-square volts per unit bandwidth) at frequency f with current I flowing, ω is $2\pi f$ and τ is some characteristic time of the postulated noise mechanism. Various physical significances have been attached to τ, such as the free life of the conduction electrons between release and capture,[14] a time constant of fluctuations in the total number of current carriers,[4] and the time of influence of stray atoms on boundary layers.[16]

If τ is single valued, this form of equation implies a spectral density independent of frequency at low frequencies, and having an inverse square variation at high frequencies. Some small intermediate range of frequencies would show an inverse first-power relationship.

Experimentally, no one seems ever to have observed low-frequency spectral density variation at any inverse power of the frequency appreciably different from unity, even at frequencies as low as 0.3 cps.[16] The present work does not include data at sufficiently low frequencies for comparison with these results, but is in agreement with the original observations of Bernamont[3] at frequencies above 25 kc in finding current-noise spectral density to vary as an inverse power of the frequency greater than 1 and approaching 2. None of the other papers listed shows data for this frequency range.

A formula of the type mentioned does not fit the results of the present work as given in Figs. 2 and 3 if τ is assumed to be single valued. It is obvious that for some distribution of values of τ such an equation could be made to fit any curve of this general shape.

The theories of current noise which have specifically included resistance value as a variable have predicted variation of mean-square noise voltage as the square of the resistance value.[16] The results shown in Fig. 4 are not incompatible with this.

The fluctuating current noise described in section III, (3) above seems not to have been described previously except for 2 passing comments in Bernamont's major publication[3] about experimental situations where his galvanometer deflections showed "mean-square fluctuations as much as 5 per cent of the mean deflection."

V. Conclusions

The results given in Figs. 3 to 6, and a great deal of additional data, lead to the following conclusions as to the relative merits of different types of resistors with respect to mean-square current noise and abnormally fluctuating current noise:

(a) The single palladium-film resistor available showed considerably less current noise than any other single resistor of comparable resistance value tested, except noninductive wire-wound units.

(b) For the same dc resistance, power rating, and conditions of measurement, the best solid carbon-composition resistors showed markedly less current noise than the best "metallized" resistors; on the other hand, many makes of solid carbon-composition resistors showed more noise than the noisiest equivalent "metallized" units.

(c) For equal power dissipation and similar measurement conditions, resistors of a given make having low power rating generally showed more current noise than those with a higher power rating; resistors *A*, *B*, and *C* in Fig. 3 are all the "metallized" type, made by 1 manufacturer, and all are dissipating the same amount of power (within 10 per cent), but have respectively power ratings of one-half, 1, and 2 watts. Whether this is a temperature phenomenon or a volume effect (as suggested by Brillouin[8] for thin metal films) has not been determined.

(d) Without exception, all records of current noise from solid carbon-composition resistors have shown abnormal fluctuations. About half the "metallized" r sistors tested have shown only normal fluctuations, co parable to those characteristic of thermal and sh effect noise. Many solid carbon-composition resist have shown fluctuations much greater than the greate obtained from " metallized" resistors.

(e) A high mean level of current noise is not necessa ily accompanied by fluctuations. Trace *E* of Fig. 5, f example, shows much greater mean level of current no than the other 4 traces, but much less fluctuation. F 6 shows, as might be expected, that for any one resist the amplitude of the fluctuations in the current no does increase as the mean noise level increases (resist *G* of Table I). Traces of current noise without abnorm fluctuations (resistor *A* of Table I) are shown for co parison. Rough calculations suggest that the amplitu of the fluctuations increases as about the fourth pow of the dc current carried by the resistor.

It can be said that, in general, large fluctuations c cur only when the mean level of the current noise high, but that there are frequent exceptions, especial with solid carbon-composition resistors.

In cases where the current noise is less than t thermal noise, as for resistor *F* in Fig. 3, the norm fluctuations of the thermal noise conceal any abnorm current-noise fluctuations that might exist.

Trace *A* of Fig. 5 shows a frequently observed ph nomenon that is probably significant. In many record the fluctuations observed during the first few minut after connecting a resistor were much greater than the rest of the record. This suggests that while the r sistance is changing to its final equilibrium value at temperature above room temperature, the change m occur in discontinuous steps, each step producing pulse of noise voltage because of the current flowing a the nature of the coupling circuits. An analogy with t Barkhausen effect in the magnetization of iron is su gested. The magnitude of the fluctuations observ must mean internal rearrangements in the resistor m terial involving units of a very large number of atom

In conclusion, it is suggested that, in many practic situations, the selection of a resistor for either low me current-noise level, or for low amplitude of curren noise fluctuations, may lead to improved operating co ditions. It seems likely that current noise may often the major component of the total noise in such equi ment as broad-band amplifiers at audio and low rad frequencies when carbon-composition type resistors a used in coupling circuits.

VI. Acknowledgments

This work has been supported, for the purchase equipment and the payment of part salary to one of (R. H. C., Jr.) by a grant from the Defence Resear Board, of the Department of National Defence, Ottaw Ont., Canada. We wish to express our appreciation f this assistance and for permission to publish the resul

An Analysis of the Sensing Method of Automatic Frequency Control for Microwave Oscillators*

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mmary—An analysis is made of the type of automatic-fre-y-control circuit that uses a simple cavity resonator for the element and either frequency modulation of the controlled tor or modulation of the cavity resonance frequency to obtain ve discriminator curves which give a null output for a fre-y equal to the average cavity resonance frequency. An analysis de of the complete automatic-frequency-control loop gain, in-g the transmission through the cavity as a function of the s parameters and the frequency-modulation swing. A discus-s presented of the best method of decreasing the pulling of the tor frequency by the cavity, and the pulling of the cavity fre-y by a load which has a variable susceptance.

Glossary

A_1 = the power attenuation between the oscillator and the cavity.

A_2 = The power attenuation between the cavity and the detector

$, b, c, d$ = the constants characterizing the general four-terminal network

$, V, n)$ = the fourier component coefficient of the frequency modulation present in the detected output

δ = per unit frequency change

$', V, n)$ = the second fourier component coefficient of the frequency-modulation frequency present in the detected output

$E_c(t)$ = the detected output voltage as a function of time

E_{c1} = the fundamental component of the sensing frequency in $E_c(t)$

E_{c2} = the second-harmonic component of the sensing frequency in $E_c(t)$

E_1 = the voltage across the input terminals of a four-terminal network

E_2 = the voltage across the output terminals of four-terminal network

f = the instantaneous frequency of the oscillator

f_0 = the resonant frequency of the cavity

ΔF = the total frequency modulation or cavity resonant frequency swing

Δf = the center frequency deviation of the oscillator from f_0

f_m = the frequency-modulation sensing frequency

G = the loop-gain, measured by breaking the loop at any point

G_a = the ratio of the output of the phase discriminator to the input of the amplifier

G_f = the ratio of the frequency deviation of the oscillator to the control-element displacement. (This may be either a voltage or a mechanical displacement depending on the control method used.)

I_1 = the current flowing into a four-terminal network

I_2 = the current flowing out of a four-terminal network

K_n = detector constant; $E_c(t) = K_n P_c^{n/2}$

n = the detector-law exponent; n is equal to unity for a linear detector, and to 2 for a square-law detector

P_{osc} = the power output of the oscillator

P_c = power incident on crystal detector

Q_1 = the input Q of the cavity

Q_2 = the output Q of the cavity

Q_0 = the unloaded Q of the cavity

Q_L = the loaded Q of the cavity

Q_{osc} = the loaded Q of the oscillator

$$\frac{\partial C(0, V, n)}{\partial W} = \frac{\partial C(W, V, p)}{\partial W}\bigg| W = 0$$

T = power relative to incident power transmitted through a cavity

$\theta = 2\pi f_m t$

$V = Q_L \Delta F / f_0$, which is the frequency-modulating swing in terms of the cavity 3-db bandwidth

$W = 2Q_L \Delta f / f_0$, which is the frequency deviation in terms of one-half the cavity 3-db bandwidth

Y_{in}, Y_{out} = admittance relative to the admittance of matched load.

Introduction

THERE ARE several methods of accomplishing automatic frequency control of oscillators operating in the microwave region. The five methods which are commonly used at present are:

1. The Pound dc discriminator.[1,2]
2. The Pound ac discriminator.[2,3]

Decimal classification: R355.6×R119.39. Original manuscript ed by the Institute, November 17, 1947; revised manuscript re-d August 19, 1948.

Electronic Research Laboratories, Air Matériel Command, ridge 39, Mass.

[1] R. V. Pound, "Frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 35, pp. 1405–1415; December, 1947.

[2] C. G. Montgomery, "Technique of Microwave Measurements, Radiation Laboratory Series No. 11," McGraw-Hill Book Co., Inc., New York, N. Y., 1947.

[3] W. G. Tuller, W. C. Galloway, and F. P. Zaffarano, "Recent developments in frequency stabilization of microwave oscillators," Proc. I.R.E., vol. 36, pp. 794–800; June, 1948.

3. The use of absorption lines.[4]
4. The method of frequency modulation.
5. The method of resonant-circuit sensing.[5]

Some of the design details of the methods (4) and (5) are to be discussed in this paper. As in all control systems, the general method of operation is as follows: The frequency of the oscillator is compared with that of the stable element (in this case a resonant circuit or a cavity resonator), with the result that a voltage (or a displacement) proportional to the amount of deviation is produced. This voltage is then applied to a frequency-controlling element on the oscillator in such a direction as to return the oscillator frequency to the frequency of the stable element. In the microwave region, a transmission cavity can be made to have the property that the power transmission is a maximum at a particular frequency, but decreases very rapidly as the frequency is deviated from that of maximum transmission. Such cavities can be made to have very high Q's and a relatively stable resonant frequency. A cavity cannot be used directly as the stable element, since the transmission is an even function of the frequency deviation from resonance. Therefore, no information can be obtained from the output, whether the deviation is plus or minus. One method of realizing a voltage which shall be positive or negative dependent upon the direction of the deviation of the oscillator frequency is to apply a small amount of frequency modulation to the oscillator frequency. It will be observed in Fig. 3 that the output of the cavity will be amplitude modulated by an amount which is dependent upon the relative amount of frequency modulation and the average oscillator frequency. In particular, it will be observed that the phase of the resulting amplitude modulation reverses with respect to the phase of the frequency modulation when the oscillator frequency passes from above the cavity resonance frequency to below it. This fact is made use of by providing an amplitude-modulation detector (usually a silicon crystal detector for the microwave region) at the output of the cavity, and a suitable amplifier to operate

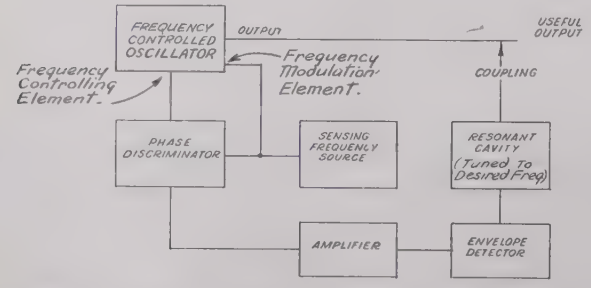


Fig. 1—Schematic of the frequency-modulating method of automatic frequency control.

[4] W. V. Smith, Jose L. Garcia de Quevedo, R. L. Carter, and W. S. Bennet, "Frequency stabilization of oscillators by spectrum lines," *Jour. Appl. Phys.*, vol. 18, p. 112; January, 1947.

[5] G. G. Gerlach, "A microwave relay communications system," *RCA Rev.*, vol. 7, pp. 560–600; December, 1946.

a phase detector. The output of the phase detector then not only proportional to the magnitude of th deviation of the average oscillator frequency from th resonant frequency of the cavity, but agrees with th algebraic sign of the deviation. Fig. 1 is a comple schematic of such a system of control. The sensing fr quency is defined as the rate of the frequency modul tion on the oscillator.

From the point of view of the over-all system b havior, it makes little difference whether the oscillat is frequency-modulated or the reasonance frequency the cavity is caused to vary about a mean frequency the sensing frequency rate. The operation is essential the same. This is the method of resonant-circuit sensin See the schematic of this system in Fig. 2. Some over-a

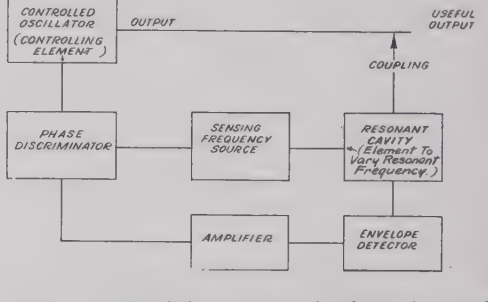


Fig. 2—Schematic of the resonant-circuit sensing method of automatic frequency control.

design considerations will dictate which method is pre erable. There are several methods available for causi the resonant frequency of a resonator or cavity to var (Since, in general, this paper will be concerned with microwave system where cavities are generally employe as the resonant system, the word cavity will be used indicate a resonator. However, the arguments and th analysis may be generally applicable to a system whic operates at a frequency where lumped components ma be used.)

1. The cavity may be equipped with a diaphragm a plate which is displaced harmonically at the sensi frequency, causing a variation of a suitable dimensio of the cavity.

2. The cavity could be constructed in such a mann that it could be resonant mechanically to the sensi frequency and driven in a suitable manner; a convenie means would be a piezoelectric crystal, a magnetostri tion bar, or even an electromechanical drive. Th changing dimensions would effect a periodic change the resonant frequency of the cavity.

3. A variable reactance could be coupled to th cavity. A nonlinear element, such as a silicon or ge manium crystal, may be used if placed in a suitab network.

In order to simplify the ensuing discussion and arg ments, only the method of frequency modulation wi be discussed. However, from the standpoint of analysi

systems are practically identical. It is only a ques- of the reference for the time scale. Hence, to apply discussion to a resonant circuit sensing system, it nly necessary to realize that the frequency swing of cavity from its resonant point is identical to the quency-modulation swing of the oscillator.

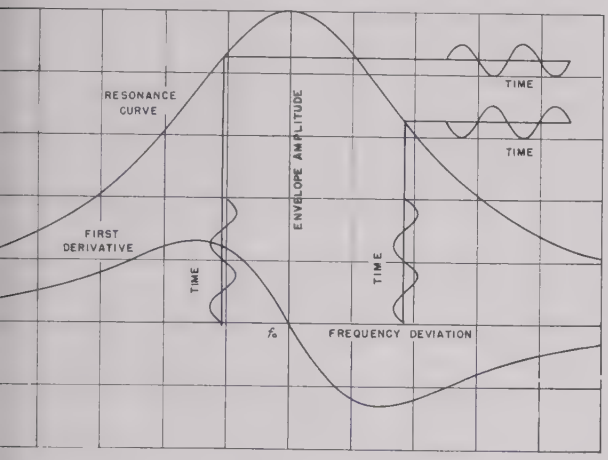


Fig. 3—Cavity characteristics.

undamental to the design of a system which uses a ity in this manner for the stable element, is the rela- ship between the frequency-modulation swing and sensitivity of the output of the phase detector in ns of the various circuit parameters. An important nent of the system is the amplitude-modulation de- or. Usually, this detector will be a crystal rectifier, ch has a law relating the input amplitude to the put amplitude not easily described. However, for -level operation the behavior will approach a square- relationship, and for high-level operation, the be- ior will approach a linear relationship. The analysis arried out for the square-law detector and the linear- detector. For the case where the operation is not cribed by either of these two laws, a heuristic ap- ximation will have to be made on the basis of experi- e and the analyzed cases. It is more convenient to uss the behavior of a cavity in terms of its Q's and nant frequency f_0 than in terms of its equivalent ped-constant parameters. The Q's and f_0 are more pted to direct measurement in the physical system. n automatic frequency-control system using the hod of frequency modulation was developed by the nor for use as the beacon-frequency automatic-fre- ncy-control system for the AN/APS-6 airborne ar.

General

he various parameters of the control-system design be taken up in term. It is assumed that the fre- ncy at which the system is to operate has been de- d. It remains, then, to decide the various other ty parameters, the loop gain of the system, and the padding that must be inserted at various locations. A cavity must be chosen which has a resonant frequency of the desired stability. The normal frequency drift of the oscillator is ascertained to determine the amount of loop gain necessary to control it within the necessary limits. Since the system is a normal negative feedback system, the frequency drift of the controlled oscillator, relative to the resonant frequency of the cavity, is equal to the frequency drift of the oscillator uncontrolled divided by one plus the loop gain. As will be seen later, the loop gain may be increased by increasing the unloaded Q of the chosen cavity, other factors remaining the same. The unloaded Q of a copper cavity is a function of the resonant frequency chosen and of its physical size. That is, a large E-mode cavity will have a much higher unloaded Q than a small re-entrant H-mode cavity for the same frequency. The physical size of the equipment may be a deciding factor in the unloaded Q obtainable.[6] For economy in the design of a control system, each component should be used to its full capability. The loop gain of the automatic-frequency-control circuit is controlled to a very large extent by the cavity parameters, and it would be wise to adjust those parameters such that the effective gain is the maximum obtainable commensurate with the other factors which must be considered.

The analysis of the loop gain in a frequency-control system with frequency modulation will be based upon three assumptions.

1. The sensing frequency and the frequency modulation swing is sufficiently low that the cavity may be considered to be in steady state for each instantaneous frequency. This will be true if the sensing frequency is small compared to the 3-db bandwidth of the cavity and if the maximum rate of change of the instantaneous frequency is less than the square of the 3-db bandwidth of the cavity. (The bandwidth must be expressed in cycles per second.)[7]

2. The relationship between the envelope amplitude applied to the detector and the output voltage can be represented by either a square-law relationship or a linear relationship.

3. The phase discriminator output is linearly related to the amplitude of the fundamental component of the detector output.

Analysis of Effective Gain Through the Cavity

The cynosure for the analysis is the cavity behavior. It is convenient to treat the high-frequency energy in terms of the incident power rather than in terms of its voltage. The reason for this is that the incident power is a relatively easy quantity to measure. Further, for work

[6] T. Moreno (editor), "Microwave Design Transmission Data," Sperry Gyroscope Co., Manhattan Bridge Plaza, Brooklyn 1, N. Y., 1944; chap. X.

[7] Balth van der Pol, "The fundamentals of frequency modulation," *Jour. IEE* (London), vol. 93, pp. 153–158; May, 1946.

of this kind, it is not necessary to know absolute impedance or admittance values, but only the relative values in the circuit. The Q's of the cavity are defined or measured under the same conditions that it will have in the operating system. Some methods of measuring the various Q's of a cavity have been described.[2,8]

It can be shown the instantaneous power transmitted through a cavity for a unit incident power is (see Appendix I)

$$T = \frac{4Q_L^2}{Q_1Q_2} \cdot \frac{1}{1 + \left[\frac{2Q_L(f - f_0)}{f_0}\right]^2}. \quad (1)$$

The instantaneous frequency of the oscillator may be expressed in terms of the sensing frequency f_m, the total frequency swing ΔF, and the deviation of the center frequency from the cavity resonant frequency and time. It is seen that

$$f = f_0 + \Delta f + \frac{\Delta F}{2} \sin 2\pi f_m t. \quad (2)$$

Since the high frequency is most easily described in terms of incident power, it is convenient to express the detector law in terms of the relationship between its incident power and the voltage output. Since it is expressed this way, it is very easy to obtain the coefficient in the expression for the limiting cases experimentally. Let the power from the cavity be fed through an attenuator to the crystal detector, which then will be considered to have an instantaneous output voltage related to the input power as follows:

$$E_c(t) = K_n P_c^{n/2}. \quad (3)$$

The combination of the previous expression and the effect of an attenuator between the cavity and the crystal detector and the effect of the decoupling between the oscillator and the cavity results in

$$E_c(t) = K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \cdot \left[1 + \left(\frac{2Q_L\Delta f}{f_0} + \frac{Q_L\Delta F}{f_0} \sin 2\pi f_m t\right)^2\right]^{-n/2}. \quad (4)$$

The above expression can be much simplified by the use of some dimensionless constants. Substituting W, V, and θ for the expressions $2Q_L\Delta f/f_0$, $Q_L\Delta F/f_0$, and $2\pi f_m t$, respectively, results in the following:

$$E_c(t) = K_n(2Q_L)^n \cdot \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \cdot [1 + (W + V \sin\theta)^2]^{-n/2}. \quad (5)$$

[8] W. Altar, "Q-circles—a means of analysis of resonant microwave systems," PROC. I.R.E., vol. 35, pp. 355–361; April, 1947.

By Fourier's theorem, the fundamental component the sensing frequency in the detected output is

$$E_{c1} = \frac{1}{\pi}\int_{-\pi}^{\pi} E_c(t) \sin\theta d\theta.$$

Combining (5) and (6), it follows that

$$E_{c1} = K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{1}{\pi}\int_{-\pi}^{\pi} [1 + (W + V \sin\theta)^2]^{-n/2} \sin\theta d\theta.$$

$$C(W, V, n) = \frac{1}{\pi}\int_{-\pi}^{\pi} [1 + (W + V \sin\theta)^2]^{-n/2} \sin\theta d\theta.$$

$C(W, V, 1)$ and $C(W, V, 2)$ are evaluated in Appen II. Figs. 4 and 5 show the plots of these functions w

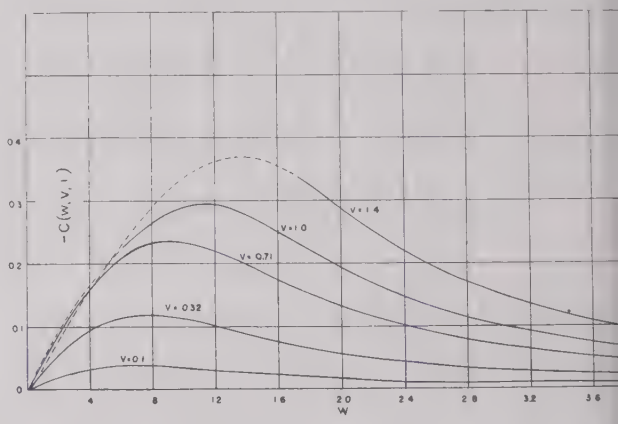


Fig. 4—Effective discriminator curves for a cavity and linear tector, W being proportional to the oscillator frequency devia while $-C(W, V, 1)$ is the relative amplitude of the dete output at the sensing frequency.

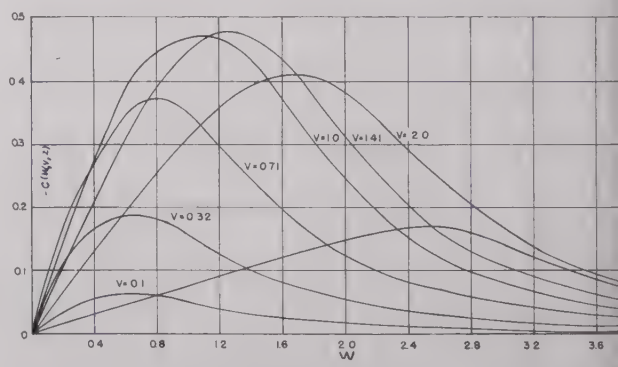


Fig. 5—Effective discriminator curves for a cavity and a square detector, W being proportional to the oscillator frequency, $-C(W, V, 2)$ is the relative amplitude of the detector outp the sensing frequency.

various values of V as a parameter, $C(W, V, n)$ i odd function of W, so it is only necessary to plot function for positive values of W.

Since, in general, the operating point of the cor system is near the W origin, the function $C(W, V$ may be approximated by the first term in its Ta expansion

$$C(W, V, n) \cong \frac{\partial C(0, V, n)}{\partial W} W, \qquad W \ll 1.$$

e response of the output to a frequency variation on input is approximately linear, and the system y be treated as a linear system. The expression $(0, V, n)/\partial W$ may be computed as a function of V. e Appendix III.) These results are plotted in Fig. 6.

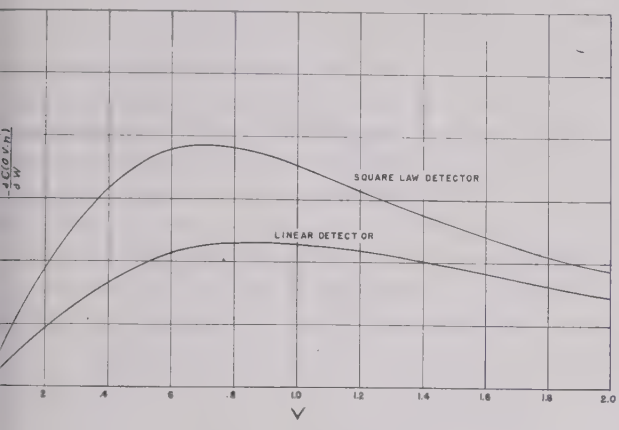


Fig. 6—Slope of the effective discriminator curves at resonance.

ese curves indicate that the function $\partial C(0, V, n)/\partial W$ s through a maximum as a function of V. For n al to unity, the maximum occurs at V equal to 0.88, for n equal to 2 the maximum occurs at V equal to 07. This indicates that in the system under design, amount of frequency modulation of the oscillator the amount of frequency swing of the resonator) be adjusted to provide a maximum gain through cavity system.

he relationship between a small frequency deviation he frequency of the power incident on the cavity and fundamental component of the sensing frequency now be completely stated and interpreted. From (8), and (9), by substitution

$$E_{c1} \cong K_n(2Q_L)^n \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{\partial C(0, V, n)}{\partial W} W, \qquad (10)$$

$$W = 2Q_L\Delta f/f_0$$

ce

$$\cong K_n(2Q_L)^{n+1} \left[\frac{P_{osc}A_1A_2}{Q_1Q_2}\right]^{n/2} \frac{\partial C(0, V, n)}{\partial W} \frac{\Delta f}{f_0}. \qquad (11)$$

as been pointed out in a preceding paragraph that $0, V, n)/\delta W$ may be maximized by the proper choice for the detector law in effect. In addition, the exsion for E_{c1} may also be maximized with the proper ce of Q relationships. As was pointed out in general ion, the unloaded Q is usually determined by the r physical factors in the system; so it remains to dis- r if there is a proper choice of the ratios of the win- or loaded Q such that the ratio of E_{c1} to Δf is a maxi- n. Differentiating the expression

$$\frac{{Q_L}^{n+1}}{(Q_1Q_2)^{n/2}} \qquad (12)$$

partially with respect to Q_1 and Q_2, equating to zero, and solving, it is seen that for maximum E_{c1} as a function of Q_1 and Q_2,

$$\text{or} \quad \left.\begin{aligned} Q_1 = Q_2 &= 2Q_0 \\ Q_0 &= 2Q_L \end{aligned}\right\} \quad \text{for} \quad n = 1, \qquad (13)$$

and

$$\text{or} \quad \left.\begin{aligned} Q_1 = Q_2 &= Q_0 \\ Q_0 &\equiv 3Q_L \end{aligned}\right\} \quad \text{for} \quad n = 2. \qquad (14)$$

While it is very interesting that such a maximum occurs, it frequently happens in the design of a system that is not desirable to take advantage of it for reasons to be considered.

Discussion of Oscillator Stability and Frequency Pulling

In coupling resonant cavities to microwave oscillators a certain amount of care must be exercised to prevent two undesirable effects:

1. If the cavity is coupled too strongly to the oscillator, the system may exhibit frequency discontinuities or jumps and refuse to oscillate at the cavity resonance frequency.

2. If the system is the type that uses the cavity-resonance frequency sensing, the varying susceptance presented to the oscillator through the coupling may cause a slight frequency and/or amplitude modulation of the oscillator at the sensing rate.

An excellent discussion of the first point appears in the literature.[1,9] Ford and Korman derived a formula which relates the frequency pulling of an oscillator to the change in load susceptance

$$\delta a = \frac{\text{(susceptance relative to matched admittance)}}{2Q_{osc}}. \qquad (15)$$

The load susceptance must, of course, be referred to the oscillator terminals. In the case of a microwave oscillator, the terminals may be defined as the point in the transmission system where the frequency of the oscillator is independent of the load conductance. The susceptance at the cavity terminals may be calculated from the formula (see Appendix I):

$$Y_{in} = \frac{Q_1}{Q_2} \cdot Y_{out} + \frac{Q_1}{Q_0} + j\frac{2Q_1\Delta f}{f_0}. \qquad (16)$$

It is observed that there is a pure susceptance change with frequency (either resonance or applied), but with a reflection through a transmission system, this susceptance variation with frequency may manifest itself as conductance variation as well as a susceptance variation at the oscillator terminals. As may be seen with the

[9] J. R. Ford and N. I. Korman, "Stability and frequency pulling of loaded unstabilized oscillators," Proc. I.R.E., vol. 34, pp. 794–799; October, 1946.

aid of a Smith chart,[10] the proper length of transmission line may favor either conductance variation or a susceptance variation. If the system under consideration uses a frequency varied resonator, the reflected susceptance variation will cause a frequency modulation of the oscillator, but the reflected conductance variation will cause an amplitude modulation of the oscillator at the sensing frequency and possibly at its harmonics. An engineering decision will have to be made as to the amount of these modulations that can be tolerated if this system is to be satisfactory. In the event that the coupling to the cavity must be reduced to bring these effects to a tolerable minimum, it may be necessary to sacrifice loop gain by the insertion of attenuation or other means to achieve the desired coupling. This point will be considered again in the discussion of the loop-gain equation.

In a similar manner to the pulling of the oscillator frequency by a susceptance presented to its terminals, the center frequency of the cavity may be pulled by a susceptance connected to its terminals. (This is one method of varying the resonance frequency.) Since, however, the cavity is to be used as the stable element of the control system, its frequency pulling must be considered, for many of the crystal detectors for microwaves have very broad tolerance on input admittance. Equation (19) may be used to calculate the pulling of the cavity by a susceptance of the detector if Q_{osc} is replaced by Q_2.

Derivation of Loop-Gain Limitations

A loop gain may be assigned to the complete loop as would be done in the case of an ordinary feedback amplifier. This is possible since E_{c1} is linearly related to the oscillator frequency deviation (11). All that is necessary to determine the loop gain is to multiply all of the individual gains together, including the various constants that relate the controlling sensitivity of the oscillator. (This last quantity may be influenced by a frequency-sensitive load upon the oscillator.[1,9]

Hence, from Fig. 1 and (11) it follows that

$$G = \frac{G_f}{f_0} G_2 K_n (2Q_L)^{n+1} \left[\frac{P_{osc} A_1 A_2}{Q_1 Q_2} \right]^{n/2} \frac{\partial C(0, V, n)}{\partial W} . \quad (17)$$

The following equation results for the gain optimized with respect to the cavity Q's and the amount of frequency modulation.

$$G = (0.23) \frac{G_f}{f_0} G_2 K_n (A_1 P_{osc})^{n/2} Q_0 . \quad (18)$$

The ratio of the frequency stabilities before and after the loop is closed can now be determined. The deviation in the oscillator frequency from the resonance frequency of the cavity after closing the loop is equal to the deviation before closing the loop divided by the quantity $(1+G)$.

[10] P. H. Smith, "A transmission line calculator," *Electronics*, vol. 12, pp. 29–33; January, 1939.

In general, it may be stated that the relation betw a small output-admittance variation and input-adm tance variation is as follows for a matched coupling s tem.

$$\left| \frac{\Delta Y_{in}}{Y_{in}} \right| = \left| \frac{\Delta Y_{out}}{Y_{out}} \right| \frac{1}{(\text{Power attenuation})}$$

where Y_{in} and Y_{out} are the input and output adm tances respectively. This expression applies to the c where the cavity is decoupled from the oscillator out by means of a directional coupler or a "magic tee" well as by a conventional attenuator. For proof of t expression, see Appendix V. It is seen, then, that the put admittance variation varies inversely with the po attenuation. In deciding the admittance variation a function of frequency due to the cavity and its ef upon the oscillator, it is observed from (19) that the mittance at the input terminals of the cavity will very large at resonance and infinite at frequencies from resonance. This is due to the fact that Y_2 is ne unity and the ratio of the input Q to the unloaded Q be chosen to be greater than unity. Hence, when u the cavity will be placed an odd multiple of quar wavelengths from the oscillator terminals (if used series system, it will be placed an integral numbe half-wavelengths from the oscillator terminals). Th fore, as seen at the oscillator terminals, a small varia of admittance will be observed as the frequency is va through the cavity resonance frequency. The magnit of this variation is seen to vary inversely with the in termination Q, since the input admittance to a quar wave section of transmission line is the reciprocal of output admittance. (In the series case, impedance is reciprocal of admittance.) Therefore, from the poin view of decreasing the oscillator pulling, is it bette increase the attenuation due to the coupling between cavity or to increase the input termination Q of the ity? The resonance frequency of the cavity is also se tive to admittance variation presented to its termi and ordinary mismatches that occur in an opera system may be responsible for some frequency variat hence, from that point of view the termination Q o cavity should be as high as possible. In looking at it is noticed that the loop gain is influenced in one t by the ratio of attenuation to Q_1; hence, this term i dependent of the decoupling for a constant amoun pulling on the oscillator. However, since the gain is portional to a power of Q_L and Q_L varies in the sam rection as Q_1 there is benefit to be reaped by increa Q_1 over that of increasing the attenuation. These s arguments will apply to the decoupling of the ca from the admittance variations of the crystal dete which might occur due to temperature variation other effects. It follows, therefore, that for the rea considered, it is preferable to decouple the cavity by use of large input and output Q's with no additiona tenuators.

Derivation of the Second-Harmonic Output of the Detector

t may be important for design consideration to know magnitude of the second harmonic of the sensing fre- ncy contained in the detector output. In some cases, component may have to be reduced to the proper plitude by suitable filtering if the amplifier and asso- ed circuits are to operate properly. In operation, the gnitude of the second-harmonic component may be ny times that of the fundamental component result- in the saturation of the amplifier. By Fourier's orem,

$$E_{c2} = \frac{1}{\pi}\int_{-\pi}^{\pi} E_c(t)\cos 2\theta d\theta \tag{20}$$

$$= K_n(2Q_L)^n\left[\frac{P_{oso}A_1A_2}{Q_1Q_L}\right]^{n/2} D(W, V, n) \tag{21}$$

re

$$(W, V, n) = \frac{1}{\pi}\int_{-\pi}^{\pi} \frac{\cos 2\theta d\theta}{[1 + (W + V\sin\theta)^2]^{n/2}} \cdot \tag{22}$$

, V, n) is evaluated in Appendix IV. Fig. 7 shows it function of V for n equal to 1 and 2. The values of

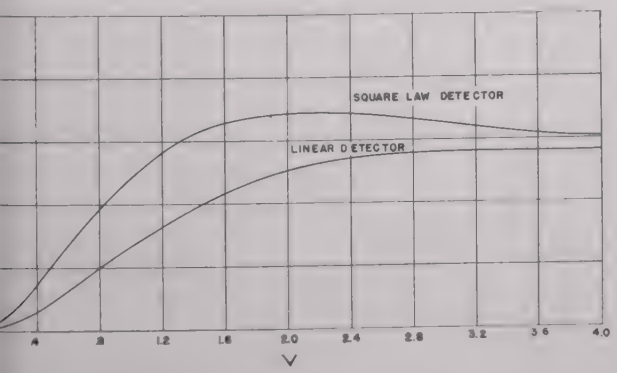


Fig. 7—Quantity of second harmonic of the sensing frequency in the detected output.

V, V, n) are not important as a function of W since operating point is near W equal zero. The value of second-harmonic component is maximum at W al zero.

Appendix I

If it is assumed that the cavity has only a single mode within the desired range of operation and has a bandwidth (to the 3-db points) which is small compared to its center frequency, then a number of simplifying assumptions can be made to expedite the analysis of its behavior. For the purposes of this paper, the cavity may be described sufficiently by the constants Q_1, Q_2, Q_0, and f_0. The restrictions and assumptions of the lumped-circuit model which will be assumed are described.[9] This paper is generally concerned with cavities that have only one window or loop, but the arguments are applicable to transmission cavities as well. Fig. 8 is a schematic of the model cavity. Use is made of a similar model in describing the behavior of cavities used as wavemeters;[11] the expressions used have a slightly different notation. The following are the definitions of the Q's:

$$Q_0 = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Power dissipated in the walls of the cavity)}}; \tag{23}$$

$$Q_1 = 2\pi f_0 \frac{\text{(Maximim energy stored in the cavity)}}{\text{(Power transmitted back through the input window)}}; \tag{24}$$

$$Q_2 = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Power transmitted through output window)}}; \tag{25}$$

$$Q_L = 2\pi f_0 \frac{\text{(Maximum energy stored in the cavity)}}{\text{(Total power lost to the cavity)}}; \tag{26}$$

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_1} + \frac{1}{Q_2}. \tag{27}$$

Since the actual values of the admittance of the load and the source to which the cavity is connected are not known, the effective shunt resistance cannot be known. However, for the purposes of this paper, it is not necessary to know this; it is only necessary to know the admittance relative to the value for which the proper Q was determined.

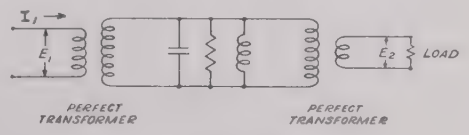


Fig. 8—Equivalent circuit of a cavity.

Since the cavity will, in general, be rather loosely coupled to the oscillator, the most useful way of describing the power transmission will be in terms of the ratio of incident power to the output power. Taking account of the reflected power at the input terminals of the cavity, by conventional circuit analysis[12] there results[13]

[11] See chap. 5 of footnote reference 2.
[12] E. A. Guillemin, "Communications Networks," John Wiley and Sons, New York, N. Y., 1931 and 1935.
[13] These expressions may be obtained directly from page 65 and page 291 of footnote reference 10 with the corresponding change in notation.

$$Y_1 = \frac{Q_1}{Q_2} Y_2 + \frac{Q_1}{Q_0} + j2Q_1 \frac{\Delta f}{f_0} \tag{28}$$

and

$$T = \frac{(2Q_L)^2}{Q_1 Q_L} \cdot \frac{1}{1 + \frac{(2Q_L \Delta f)^2}{f_0^2}} \cdot \tag{29}$$

Appendix II

For n equal to unity, the following integral is to be evaluated

$$C(W, V, 1) = \frac{1}{\pi} \int_{-\pi}^{\pi} \frac{\sin \theta d\theta}{\sqrt{1 + (W + V \sin \theta)^2}} \cdot \tag{30}$$

From a practical point of view, it is most expedient to evaluate this by a series expansion about V equal to zero. Expanding the denominator by a Taylor series, then, it is seen that

$$C(W, V, 1) = \frac{1}{\pi} \int_{-\pi}^{\pi} \sum_{n=0}^{\infty} \frac{d^m}{dW^m} \frac{V^m \sin^{m+1} \theta d\theta}{n! \sqrt{1 + W^2}} \cdot \tag{31}$$

Exchanging the order of integration and summation and integrating,[14]

$$C(W, V, 1) = 2 \sum_{m=0}^{\infty} \frac{1 \cdot 3 \cdots (2m+1)}{2 \cdot 4 \cdots (2m+2)} \frac{d^{2m+1}}{dW^{2m+1}} \frac{1}{\sqrt{1 + W^2}} \frac{V^{2m+1}}{(2m+1)!} \cdot$$

Performing the differentiating and factoring gives the required series

$$C(W, V, 1) = \frac{-WV}{(1 + W^2)3/2} \left[1 + \frac{3(2W^2 - 3)}{8(1 + W^2)^2} V^2 + \cdots \right.$$
$$+ \frac{5(8W^4 - 40W^2 + 15)}{64(1 + W^2)^4} V^4$$
$$\left. + \frac{35(16W^6 - 168W^4 + 210W^2 - 35)}{81{,}920(1 + W^2)^6} V^6 \cdots \right]. \tag{33}$$

This series is convergent for V^2 less than $1 + W^2$ since the radius of the circle of convergence of the series for $1/1 + (W+z)^2$ in the Z-plane is $\sqrt{1+W^2}$, when expanded about $Z = 0$.

[14] Mathematical Tables from "Handbook of Chemistry and Physics," Chemical Rubber Publishing Co., Cleveland, Ohio, 5th edition, 1936; no. 337, p. 271.

For n equal to 2, the following integral must be evaluated

$$C(W, V, 2) = \frac{1}{\pi} \int_{-\pi}^{\pi} \frac{\sin \theta d\theta}{1 + (W + V \sin \theta)^2} \cdot$$

By expanding the integrand in partial fraction, it is shown that[15]

$$C(W, V, 2) = \frac{1}{2V\pi j} \int_{-\pi}^{\pi} \left[\frac{W + j}{W + j + V \sin \theta} - \frac{W - j}{W - j + V \sin \theta} \right] d\theta.$$

$$C(W, V, 2) = \frac{1}{Vj} \left[\frac{1}{\sqrt{1 - \left(\frac{V}{W + j}\right)^2}} - \frac{1}{\sqrt{1 - \left(\frac{V}{W - j}\right)^2}} \right].$$

By a few algebraic simplifications, it is found that

$$C(W, V, 2) = -\frac{2\sqrt{W^2 + 1}}{V \sqrt{(W^2 + 1)^2 - 2V^2(W^2 - 1) + V^4}} \cdot \sin \tfrac{1}{2} \arctan \frac{2V^2 W}{(W^2 + 1)^2 - V^2(W^2 - 1)} \cdot$$

Appendix III

The functions representing the slopes of the effect discriminator curves at the origin are to be evaluat The slope of the curves for the linear detector may evaluated rather easily in terms of complete elliptic in grals from the original integral.

$$\frac{\partial C(0, V, 1)}{\partial W} = -\frac{V}{\pi} \int_{-\pi}^{\pi} \frac{\sin^2 \theta d\theta}{(1 + V^2 \sin^2 \theta)^{3/2}} \cdot$$

Substituting $1 - \cos^2 \theta$ for $\sin^2 \theta$ and $\theta = \pi/2 - \phi$, it lows that

$$\frac{\partial C(0, V, 1)}{\partial W} = \frac{-4V}{\pi(1 + V^2)^{3/2}} \int_0^{\pi/2} \frac{(1 - \sin^2 \phi) d\phi}{\left[1 - \frac{V^2}{1 + V^2} \sin^2 \phi\right]^{3/2}} \cdot$$

Consulting de Haan,[16] and making a few algebraic s plifications, it is seen that

[15] See no. 349, p. 272, of footnote reference 14.

[16] D. Bierens de Haan, "Nouvelles Tables D'integrales Defi G. E. Strechert and Co., New York, N. Y.; 1939; p. 91.

$$\frac{\partial C(0, V, 1)}{\partial W} = \frac{-4}{\pi V\sqrt{1+V^2}}\left[K\left(\frac{V}{\sqrt{1+V^2}}\right) - E\left(\frac{V}{\sqrt{1+V^2}}\right)\right] \quad (40)$$

re K and E are elliptic integrals.[17]

he slopes of the effective discriminator curves for the re-law detector may be found most easily by dif-ntiating the function directly, resulting in

$$\frac{\partial C(0, V, 2)}{\partial W} = \frac{-2V}{(1+V^2)^{3/2}}. \quad (41)$$

Appendix IV

he integral representing the amount of second har-ic of the sensing frequency may be worked out at W l to 0. For the linear detector,

$$D(0, V, 1) = \frac{1}{\pi}\int_{-\pi}^{\pi} \frac{\cos 2\theta d\theta}{\sqrt{1+V^2\sin^2\theta}}. \quad (42)$$

stituting $1-\cos^2\theta$ for $\sin^2\theta$, $n/2-\phi$ for θ, $1-2\sin^2\phi$ cos 2θ, and shifting the integration intervals, it is that

$$V, 1) = \frac{-4}{\pi\sqrt{1+V^2}}\int_0^{\pi/2} \frac{(1-2\sin^2\phi)d\phi}{\sqrt{1-\frac{V^2}{1+V^2}\sin^2\phi}}. \quad (43)$$

found that[18]

$$, V, 1) = \frac{-4}{x\sqrt{1+V^2}}\left[K\left(\frac{V}{\sqrt{1+V^2}}\right) - 2D\left(\frac{V}{\sqrt{1+V^2}}\right)\right]. \quad (44)$$

the square-law detector,

$$D(0, V, 2) = \frac{1}{\pi}\int_{-\pi}^{\pi} \frac{\cos 2\theta d\theta}{1+V^2\sin^2\theta}. \quad (45)$$

stituting $\frac{1}{2}(1-\cos 2\theta)$ for $\sin^2\theta$, $\phi=\theta/2$, and shifting nterval of integration, it is seen that

$$V, 2) = \frac{2}{\pi(2+V^2)}\int_0^{2\pi} \frac{\cos\phi d\phi}{1-\frac{V^2}{2+V^2}\cos\phi}\, d\phi. \quad (46)$$

king some algebraic simplifications,[19] it is found that

$$D(0, V, 2) = \frac{4}{V^2}\left[\frac{2+V^2}{2\sqrt{1+V^2}} - 1\right]. \quad (47)$$

Appendix V

The expression for the relation between the variation in input admittance and the variation in output admittance for a matched four-terminal network may be derived easily using matrix methods.[12] Consider the general four-terminal network characterized by the four constants a, b, c, and d. Not all are independent, for a bilateral system with a consistent set of units $ad-bc$ must equal unity. Consider the network operating into and admittance Y_{out} with an associated input admittance Y_{in}. Let E_1 and I_1 be the input voltage and current to the network, and let E_2 and I_2 be the output voltage and current. It follows, then, that

$$\begin{vmatrix} E_1 \\ I_1 \end{vmatrix} = \begin{vmatrix} a & b \\ c & d \end{vmatrix} \times \begin{vmatrix} 1 & 0 \\ Y_{out} & 1 \end{vmatrix} \begin{vmatrix} E_2 \\ 0 \end{vmatrix}.$$

Hence,

$$E_1 = (a + bY_2)E_2$$

and

$$I_1 = (c + dY_2)E_2.$$

By dividing I_1 by E_1, it follows that

$$Y_{in} = \frac{c + dY_{out}}{A + BY_{out}}.$$

By differentiating with respect to Y_{out}, it follows that

$$\Delta Y_{in} \cong \frac{\Delta Y_{out}}{(a + bY_{out})^2}.$$

The power attenutaion for the network can be derived for the relationship between E_1 and E_2 if Y_{in} and Y_{out} be considered as pure conductances. If they are not, then the power input and output expression must contain only the real parts of Y_{in} and Y_{out}. To include this refinement generalizes the final expression, but it does not increase its usefulness. Hence,

$$\text{(Power attenuation)} = \left|\frac{E_2^2\ Y_{out}}{E_1^2\ Y_{in}}\right| = |a + bY_{out}|^2 \frac{Y_{out}}{Y_{in}}.$$

It then follows that

$$\left|\frac{\Delta Y_{in}}{Y_{in}}\right| = \left|\frac{\Delta Y_{out}}{Y_{out}}\right| \frac{1}{\text{(Power attenuation)}}.$$

Even though the admittances Y_{in} and Y_{out} must be real, the variations do not have to be real and, in general, they will not be real. If it were possible to find the values of a, and b and Y_{out} for the particular network, it would be possible to calculate the relationship between the real and imaginary components of the admittance variation. However, if only the power attenuation is known, only the relation between the absolute magnitudes of $\Delta Y_{in}/Y_{in}$ and $\Delta Y_{out}/Y_{out}$ can be found.

E. Jahnke and F. Emde, "Tables of Functions with Formulae Curves," Dover Publications, New York, N. Y., 1943.
See p. 73 of footnote reference 17.
See p. 272, no. 349, of footnote reference 14.

New Design for a Secondary-Emission Trigger Tube

C. F. MILLER†, SENIOR MEMBER, IRE, AND W. S. McLEAN‡

Summary—**Aspects of the design of a nongaseous miniature trigger tube are given. A triode input section produces a primary electron beam which impinges on a dynode to produce secondary electrons. These are collected by two different output elements which may be used separately or as a unit. A surface having long life and stability is described. Suggested applications include its use as a relaxation oscillator, multivibrator, pulse inverter, triangular-wave generator, and dynatron.**

Introduction

AN EARLIER PAPER[1] described the application of secondary-emission phenomena to the problem of obtaining trigger action. A design of tube was shown which served well for the basic principles involved, but which left something to be desired in terms of high-frequency operation and miniaturization. The tube described in this present paper satisfies these last two requirements. It carries the designation NU-1032-J.

Significance of the "First-Zero" Voltage

It has become common practice, in working with secondary-emission surfaces or dynodes, to designate the quality by the value of the "first-zero" voltage taken from the dynatron characteristic. Such a characteristic is shown as Fig. 1, curve A. The "first-zero" voltage is the lower of two voltages at which the current to the dynode is zero. For the construction used, at this voltage there will be a large difference of potential between the collector grid and the dynode, so that all the secondary electrons will be drawn away from it. Thus, this "first-zero" voltage is that voltage at which the ratio of secondary to primary electrons (δ) is equal to unity. It is also the voltage for the primary electrons at which their velocity is just great enough to produce one secondary electron for each primary electron.

Tube Design

The main problem which confronts the development of any type of secondary-emission tube is one of stability during life. Maintaining a low and consistent value of "first-zero" voltage is the prime requisite for a good trigger tube. In 1938, it was found[2] that contamination given off by the cathode caused poisoning of the dyn surface and consequent instability in secondary-e sion tubes. Theory and experiments showed that contamination traveled in straight lines. With thi mind, the first stable secondary-emission amplifier developed by sheilding the dynode from the cathode causing the primary electrons to travel in curved pa

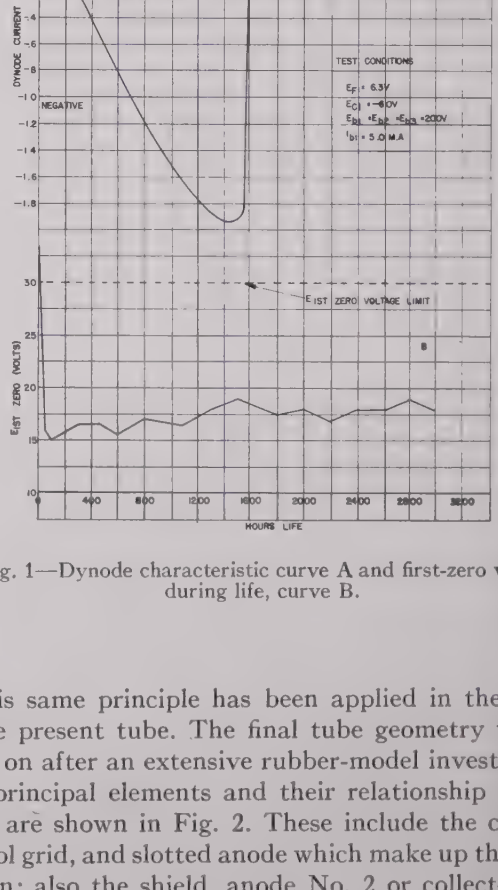


Fig. 1—Dynode characteristic curve A and first-zero value during life, curve B.

This same principle has been applied in the des of the present tube. The final tube geometry was cided on after an extensive rubber-model investigat The principal elements and their relationship to e other are shown in Fig. 2. These include the catho control grid, and slotted anode which make up the tri section; also the shield, anode No. 2 or collector g dynode, and anode No. 3. The slot in the plate form beam of primary electrons. The influence of the shi which is at cathode potential, causes the primary e trons to travel in curved paths, passing through an No. 2 and finally striking the dynode. Particular was paid to obtaining an even density and spread electrons over the surface of the dynode in order prevent overloading any one portion of the surfa Electrons passing through the middle of the slot str

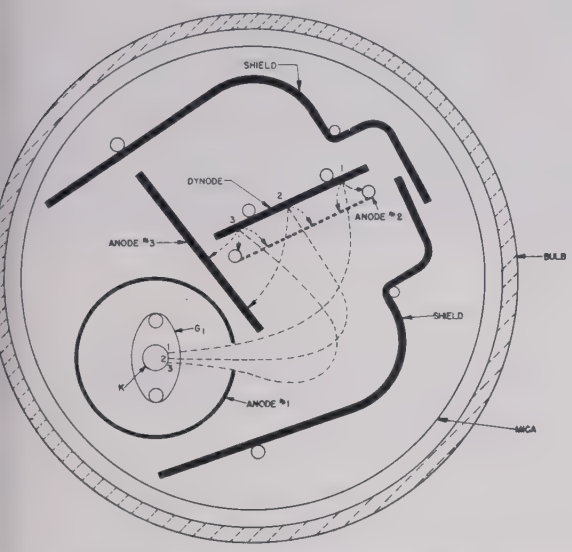
* Decimal classification: R339. Original manuscript received by the Institute, April 16, 1948; revised manuscript received, February 3, 1949. Presented, 1948 IRE National Convention, New York, N. Y., March 25, 1948.

† National Union Radio Corp., Orange, N. J.
‡ Western Electric Co., New York, N. Y.

[1] A. M. Skellett, "Use of secondary electron emission to obtain trigger or relay action," *Jour. Appl. Phys.*, vol. 13, pp. 519–524; August, 1942.

[2] J. L. H. Jonker and A. J. W. M. v. Overbeck, "The applications of secondary emission in amplifying valves," *Wireless Eng.*, vol. 15, pp. 150–157; March, 1938.

middle of the dynode, and those through the outer ts spread evenly to each side of the center.
he rubber-model trajectories were checked by means tube containing a fluorescent dynode which showed same spread of electrons on the dynode as did the



g. 2—Cross section showing the elements of the trigger tube.

er model. One of the innovations on this type of is the use of anode No. 3. Previous designs de- led entirely on the collector grid for collecting the

Fig. 3—Photograph of the miniature trigger tube.

ndaries, and, consequently, many secondaries in ing through the collector grid struck the first anode. de No. 3 prevents this loss in that it shields the first anode almost completely from secondaries passing through the collector grid and eliminates the conditions which were conducive to oscillation by the secondaries about the collector grid.

The dynode surface combines a base metal of oxygen-free copper or silver with a thin composite layer of barium and magnesium oxides. The barium and magnesium are evaporated onto the copper under vacuum conditions. The base metal is highly polished on the active side before the layers of the other metals are applied. Special care is taken not to contaminate the copper in any way once it is polished. The thin layers of barium and magnesium are oxidized by exposing the structure to oxygen. It is believed that some of the pure metal diffuses through this oxide layer to the surface during the aging process.

A photograph of the tube is shown in Fig. 3.

Life Tests

The arbitrary limit for an acceptable value of secondary-emission ratio for a trigger tube is a "first zero" of +30 volts or less. Curves A and B in Fig. 1 show what may be expected of the tube with respect to life and consistency; curve A is the dynode characteristic taken after 3,000 hours life. Curve B shows that the "first-zero" value has been good throughout life. The initial drop (from approximately 35 to 16 volts) indicates that the surface was "aged-in" during this time, and after a few hours the "first zero" reached an average value varying between 15 and 19 volts. The tests indicate that this particular surface will maintain consistent values of δ under 0.5 watt per square centimeter dissipation.

These life tests have shown that it is possible to make dynodes of the kind used herein that last for 5,000 hours or better, maintaining constant characteristics.

Applications

The basic trigger circuits for the use of this tube are described in the earlier paper[1] referred to above.

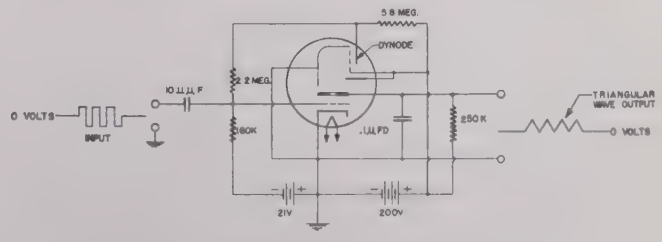


Fig. 4—Triangular-wave generator, $f = 5{,}000$ cps.

Fig. 4 shows the tube used as a triangular-wave generator. The action of the circuit is as follows: The square-wave input is differentiated into triggering

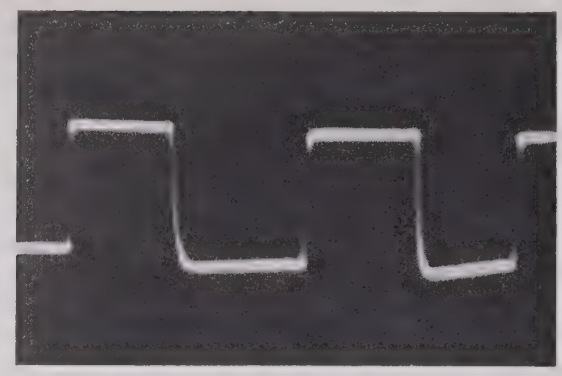


Fig. 5—Square-wave output for 100 kc.

pulses by the input *RC* circuit. These pulses trigger tube "on" and "off," as in the circuit described earl When the tube is triggered "off," the 0.1-μf capacito charged exponentially through the plate resis (250,000 ohms). When the tube is triggered "on," output capacitor discharges exponentially through tube.

Fig. 5 shows the output wave form when the tub triggered at 100 kc. The leading edge has a rise time approximately 0.2 microsecond. The full capabilities this tube have not yet been determined; however, i believed that it can be made to trigger at frequencies high as 1 Mc since no gas-deionization time is involv

Electronic Techniques Applied to Analogue Methods of Computation*

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Summary—**This paper describes in detail the electronic devices and principles that have been developed for the California Institute of Technology (CIT) electric analogue computer. This is a general-purpose, large-scale computer applicable to a wide range of linear and nonlinear ordinary algebraic or differential equations and linear and nonlinear partial differential equations.**

In addition to the basic principles of the computer, a detailed discussion is given of those elements considered to be of particular interest. These include the devices for generating the arbitrary functions of the independent variable (the excitation functions), the amplifiers for producing active linear elements such as negative impedances and for representing the nonsymmetrical terms of the matrix specifying the differential equations, the multipliers for producing arbitrary functions of the dependent variables (nonlinear elements).

Performance data on these devices are presented, together with analogies and solutions of representative types of problems.

INTRODUCTION

THE ELECTRIC analogue computer, which has been developed by the California Institute of Technology provides a device having as wide a field of application as is considered practical with the electric analogue principle. It is thought that the greatest field of usefulness of such a computer lies in the accuracy range that is limited to the order of 1 per cent, which covers the vast majority of problems in the general field of engineering analysis. With this accuracy-limitation, great simplification in the computer and its operation results from the use wherever possible of *R*, *L*, *C*, and transformer circuits in simulating the linear terms of both algebraic and differential equations This is in contrast to computers which employ electro amplifiers for all integration or differentiation. Th provide higher accuracy, but are restricted in their fi of application as compared to that practicable with computer of the type discussed here.

Electric analogue computers are limited to one in pendent variable that can be continuously represent However, by the use of finite-difference methods, su as must be employed for digital computers, more in pendent variables can be handled. The CIT compute designed to handle partial differential equations w up to three independent variables. Two basic circ techniques are employed. For the first of these, one dependent variable is continuously represented as ti on the computer.[1,2] In the other, all independent va ables are represented by finite-difference methods, which case the analogous circuit becomes a steady-st ac or dc mesh.[3] Both techniques are applicable to b ordinary and partial differential equations. It is the a plication of the first technique that requires most of t electronic equipment, such as devices to generate sui ble functions of the independent and dependent va ables, multipliers for multiplying any two variables gether, and amplifiers to provide the proper pow output levels from these devices and to produce ne tive impedance terms and the unilateral or unsymme cal terms of a matrix or servosystem.

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† California Institute of Technology, Pasadena, Calif.

[1] G. D. McCann, "The mechanical transients analyzer," *P NEC*, vol. 2, p. 372, 1946.

[2] E. L. Harder and G. D. McCann, "A large scale general-purp electric analog computer," AIEE Technical Paper 48-112.

[3] G. D. McCann and C. H. Wilts, "Application of electric ana computers," *Jour. Appl. Mech.*, 1949.

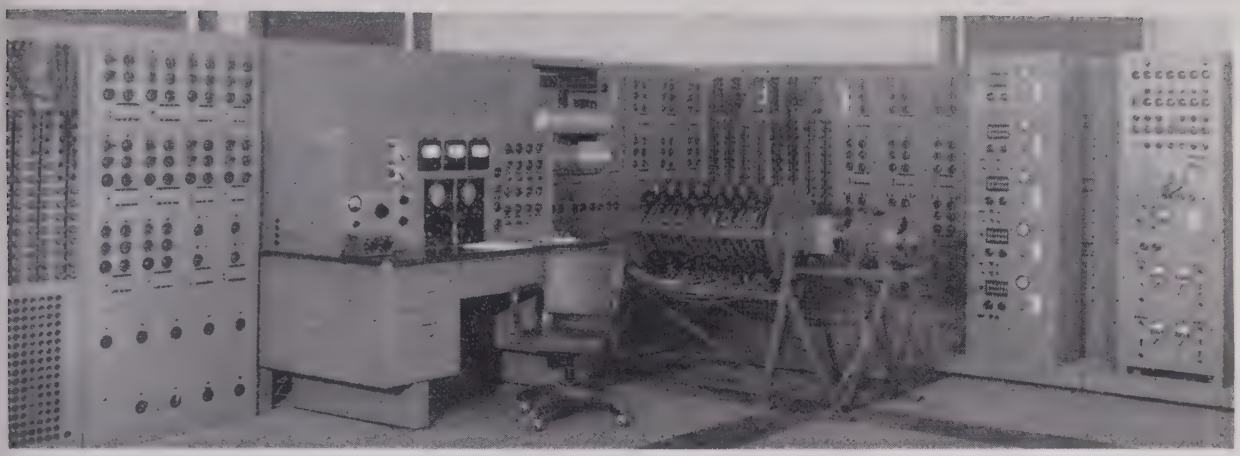


Fig. 1—A general view of the California Institute of Technology computer.

t is these recently developed basic electronic ele-
nts of the computer that are to be described in this
per. The general arrangement of the computer, a
otograph of which is shown in Fig. 1, has been de-
bed in a previous paper.[2]

Examples of Computer Techniques

The required specifications for the analogue elements
best considered by first illustrating their application
specific problems and describing the analogue tech-
ues employed. The basic analogue elements em-
yed in the computer are listed in Fig. 2 in terms of
ir symbolization as employed for circuit layouts.
pical analogous circuits for ordinary differential
ations are given in Figs. 3 and 4.

As illustrated in the examples, R, L, and C circuit ele-
nts are usually employed for all linear "self-imped-
e" terms. They are also employed for bilateral mu-
l terms that can be formed with junction points be-
en the meshes simulating the simultaneous equations
for transfer functions. Special transformers[2] are
d wherever possible for bilateral mutuals requiring
uit isolation or, in conjunction with amplifiers, to
vide a four-terminal unilateral element.

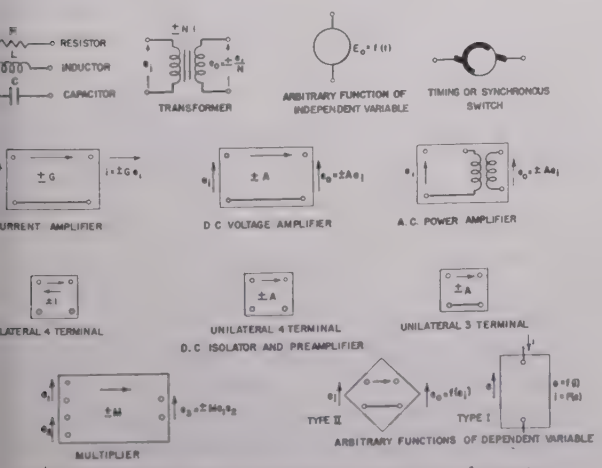


Fig. 2—Symbolization of the basic computer elements.

The use of such circuit elements fixes the impedance level and the rate of computation. To minimize stray coupling and stray pickup problems and enable the quick setup of a wide range of circuits, the circuit impedance base must be kept low and the power level reasonably high. Unintentional impedances, such as the output impedance of a voltage amplifier, must be kept below a few ohms if they are series elements, and above several hundred thousand ohms (and in extreme cases, 10 or 20 megohms) if they are shunt impedances. In order that the elements are sufficiently close to being ideal inductors, etc., the natural frequencies of the analogue circuits must lie within certain limits. A normal operating frequency range of from 10 cps to 1 or 2 kc was chosen.[2] However, because of this wide frequency range, transformers have certain limitations. It is difficult to keep the ratio of the series to magnetizing impedance of a transformer sufficiently low over such a wide operating frequency range. This, together with

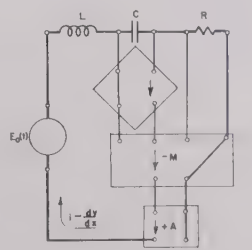


Fig. 3—Illustrative analogy for an ordinary differential equation:

$$F_0(x) = k_1 \frac{d^2y}{dx^2} + f_1(y)\frac{dy}{dx} + k_2 y.$$

Analogous circuit equation:

$$E_0(t) = L\frac{d^2q}{dt'^2} + \frac{AMR}{C}F_1(q)\cdot\frac{dq}{dt'} + \frac{q}{C}\left\{R\frac{dq}{dt'}\ \text{negligible}\right\}.$$

Numerical conversion formula:

$$y = a\frac{F_0}{E_0}q; \quad L = \frac{a}{n^2}k_1; \quad \frac{I}{C} = ak_2; \quad \frac{AMR}{C}\cdot F_1(q) = \frac{a}{n}f_1(q)$$

where
a = impedance-base change factor
n = independent-variable conversion factor.
(One unit of x in its system of units equals $1/n$ units of t' seconds.

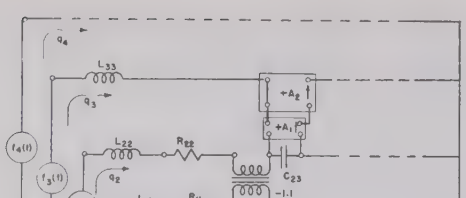


Fig. 4—Illustrative analogy for simultaneous equations:

$$f_1(t) = K_{11}\frac{d^2y_1}{dt^2} + k_{11}\frac{dy_1}{dt} + k_{12}\frac{dy_2}{dt} + \cdots$$

$$f_2(t) = K_{22}\frac{d^2y_2}{dt^2} + k_{22}\frac{dy_2}{dt} + k_{12}\frac{dy_1}{dt} + \cdots$$

$$f_3(t) = K_{33}\frac{d^2y_3}{dt^2} + k_{23}y_2 + \cdots$$

$$f_4(t) = \cdots .$$

the tendency of transformer-coupled feedback circuits to "motorboat," constitute the major difficulties sometimes encountered in the use of transformers. It is, therefore, occasionally necessary to use four-terminal "isolators" which operate properly from zero frequency (dc) to the upper limit of 1 or 2 kc (see Fig. 4). The general philosophy regarding accuracy has been to achieve 1 per cent accuracy in all elements. Thus, an inductor should exhibit a variation in effective inductance which is less than 1 per cent of its value throughout the range in operating level and operating frequency (in this case, 100 to 1,000 cps). In addition, its effective resistance should be less than 1 per cent of its reactance at all frequencies and levels (i.e., its Q should be greater than 100). Similarly, all of the electronic equipment must have less than 1 per cent amplitude distortion throughout the operating range. Furthermore, the frequency response of the equipment must not vary more than 0.1 db throughout the frequency range of the computer. It was considered satisfactory if the phase shift were less than 1° throughout the frequency range. However, since these devices are frequently used in cascade and in feedback circuits, the phase and amplitude characteristics must be controlled to eliminate oscillations at high and low frequencies. To prevent low-frequency "motorboating," all electronic equipment except the arbitrary function of the independent-variable (forcing-function) generators was designed for operation from zero frequency (dc) to the upper frequency limit of the computer. To simplify elimination of high-frequency oscillations, it was found desirable to extend the high-frequency limit of all equipment as far as practicable.

Power Amplifiers

Three types of amplifiers are shown in Fig. 2—ac power amplifiers, dc current generators, and dc voltage amplifiers. Although they are frequently used for other purposes, their most common uses are: (1) Ac power amplifier: to amplify the output of the forcing-function generators (arbitrary function of the independent variable) and provide a low source impedance. (2) Dc c rent generator: to produce a current as a "forcing fu tion." (3) Dc voltage amplifier—to produce negative unilateral impedances, and to amplify the output generators of an arbitrary function of the depend variable. Typical examples of the use of the dc am fiers are given in Figs. 3 and 4. The dc amplifier in impedances must be kept high so that they will not c tort the circuit across which they are to be connect However, since they usually can be used with a sm input signal, the circuits can be so set up that 500,0 ohms is a sufficiently high input impedance. For m problems, power levels corresponding to a maximu current of 30 to 50 ma and a voltage of 50 to 100 vo are adequate.

Two types of dc voltage amplifiers were found desi ble. One of these has a negative or 180° phase-rever characteristic, and is to be used particularly for ser mechanism problems. It has a maximum gain of 50 a is usually used in groups of three or more. The other i positive-gain amplifier. For this a gain of 100 is usua adequate when it is used to produce a negative or u lateral impedance, since the accuracy limitation of t computer is 1 per cent. However, in limited cases, larg gains are required. In order to provide for this ad tional gain, the amplifier can be used with a preamp fier, which has a gain variable from 0 to 10. It is co sidered adequate if the current amplifiers can deliv about 50 ma and 100 volts. Their effective source i pedance should be a megohm or more.

As can be seen from the illustrative circuits (Figs. and 4), it is necessary that all analogue elements capable of operation with no connection between no mal ground terminal and true system ground, or th must be provided with isolating elements at their te minals. As will be discussed later, it was found pra ticable to design the electronic equipment for operati with no connection to system ground, thereby affordi considerable simplification.

Multipliers

The multipliers must be capable of multiplying t gether any two variable voltages having frequencies u ti 1 or 2 kc. Atypical examples of their use is given Fig. 3. Recalling the over-all 1 per cent accuracy r quirements of the computer, it is desirable, as a *min mum* requirement, that the multipliers be capable forming products of functions such that the instantan ous product obtained for any input voltages deviat from the correct value by an amount less than 1 p cent of the maximum product obtainable.

Arbitrary Function Devices

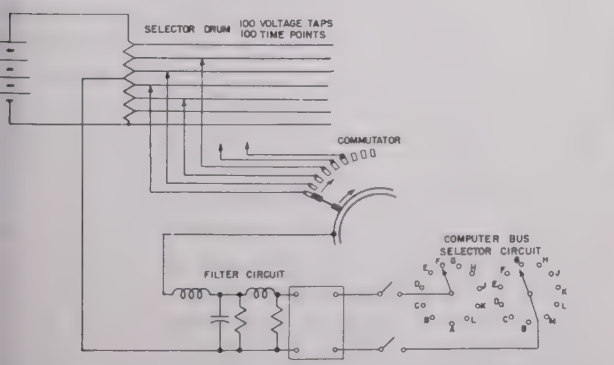
Because of the wide range of possible functions, it desirable to have several methods of generating fun tions of both time, the independent variable, and t dependent variables. For the "independent variable" time functions, six square-wave voltage generators, co

ing of storage-battery circuits, and six variable-fre-ncy sine-wave generators are provided. These are cribed in footnote reference 2. In addition, however, s necessary to be able to generate any time function t can be plotted or expressed mathematically. Many portant nonlinear functions of the dependent vari-e, such as saturation curves, square-law character-cs, sudden discontinuities, etc., can be generated with tively simple diode circuits. In addition, however, vision must be made for perfectly arbitrary func-s.

Description of Electronic Components

erator of Arbitrary Functions of the Independent Variable

ince the computer operates in the audio-frequency ge, it is natural to consider the possible application he various principles employed for sound recording, uding the light-beam–photocell principle, magnetic e, or wire recording and conventional disk recording playback techniques. All of these methods were l experimentally and found practical. As will be ussed later, the light-beam–photocell principle could most easily and practically developed into a form able to the needs of the computer, and was adopted one form of "forcing-function" generator. However, use of any of the above principles leads to one prac-l difficulty. If the function of the independent vari- is unknown and the only known data are some al solutions obtained from experimental tests, it is rable to have a generator in which the arbitrary tion can be varied until, by trial and error, a cor- solution is obtained.



5—Schematic diagram of the arbitrary forcing-function genera-or Type I; variable gain amplifier, 1-ohm output impedance, 50 olts at 1-ampere output.

e I

he device illustrated in Fig. 5, although limited in maximum frequency which it can represent, does vide for rapid changes in the forcing function. As wn in the schematic diagram (Fig. 5), 100 voltage s are provided which may be selected by the adjust-t of 100 switches, each of which connect to a seg-t of a commutator, thus providing 100 points in e. A filter circuit provides for a smooth transition between points. The 100 voltage taps are rods placed around the periphery of a circular drum, over which are 100 circular rings each connecting to a segment of a mechanical commutator. Adjustment of the function generator is accomplished by rotating the rings around the drum. Three of these devices are provided in the computer. They may be used separately, or two may be ganged in series to provide up to 200 points in time.

Type II

The generator employing the photoelectric principle is illustrated in Fig. 6. A photographic film is produced

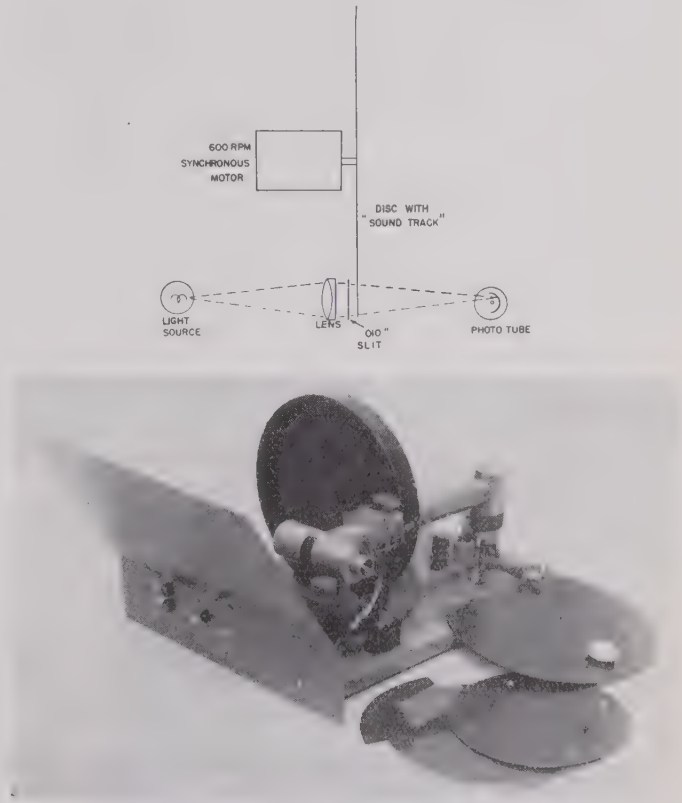


Fig. 6—(a) Schematic diagram of the generator of an arbitrary function of independent variable. (b) Photograph of a generator of an arbitrary function of independent variable.

with a circular "sound track" having a variable width proportional to the desired function. This film and a narrow slit are interposed between a light source and a phototube. When the film is rotated at synchronous speed (10 rps), a voltage is produced which is a periodic function of time, and which is, during each cycle, proportional to the desired function of the independent variable. The film diameter is 11 inches, with provision for a maximum track width of $\frac{5}{8}$ inch. A simple ac amplifier is used to increase the signal level to about 1 volt.

There are six principal sources of distortion in such an optical-photoelectric system. Two of these result from geometrical properties of the system. One arises if the width of the "sound track" is made exactly propor-

tional to the desired function. In this case, the angle subtended by the track is not exactly proportional to the desired function, but is related by a simple trigonometric expression which departs appreciably from linearity if the angle becomes too large. Since the light transmitted is proportional to the angle, a lack of linearity results. The other distortion is a frequency-sensitive amplitude distortion. It arises because of the finite slit width and limits the detail possible because of the averaging effect over the width of the slit. These distortions were made less than 0.1 per cent by maintaining suitable ratios of critical dimensions.

Four other sources of distortion are due to imperfections in the apparatus: variable sensitivity of the phototube surface, imperfect slit, variation of source light intensity with direction, and variation in transmissibility of film. Distortion arising from variable sensitivity of the phototube surface is minimized by focusing the light from the source to one spot on the phototube by means of a simple achromatic lens. Distortion due to the other three causes is believed to be the major source of error. For a perfect film, the generator shown in Fig. 6 with a $\frac{5}{8}$-inch sound track has a maximum over-all distortion of 0.3 per cent.

Power Amplifiers

Both the current amplifiers and the transformer-coupled voltage amplifiers are of conventional design and do not merit detailed discussion. For the latter, four push-pull–parallel 6L6's with approximately 60 db of negative feedback met the special output requirement of 50 volts and 1 ampere with an internal impedance of 1 ohm. The amplification varies less than 0.1 db, and th phase shift is considerably less than 1° from 10 to 2,0 cps.

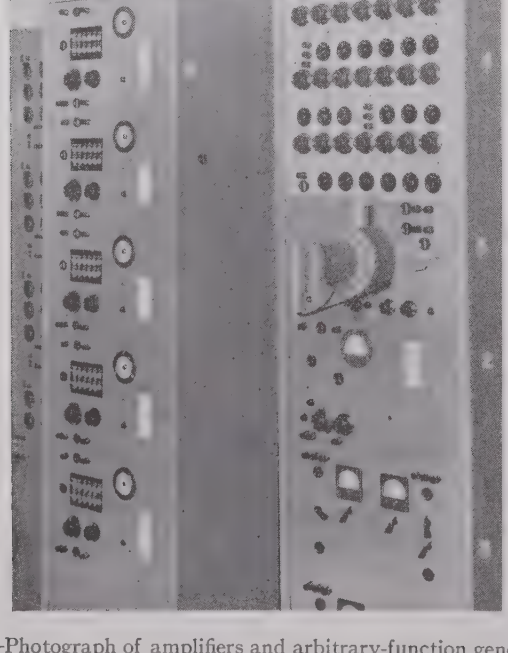


Fig. 7—Photograph of amplifiers and arbitrary-function generators. (1) Negative gain servo amplifiers. (2) Positive-gain dc amplifier. (3) Voltage and current limiters. (4) Dependent-variable function generator—Type I. (5) Dependent-variable function generator—Type II.

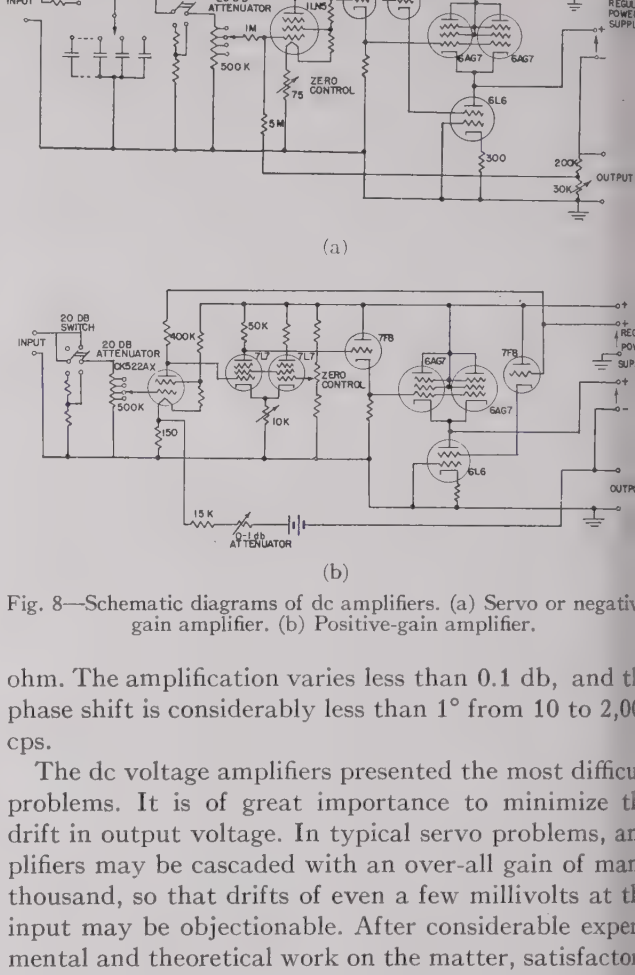
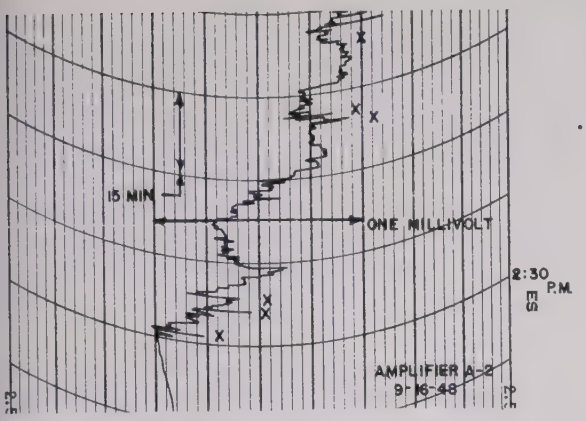


Fig. 8—Schematic diagrams of dc amplifiers. (a) Servo or negativ gain amplifier. (b) Positive-gain amplifier.

The dc voltage amplifiers presented the most difficu problems. It is of great importance to minimize th drift in output voltage. In typical servo problems, an plifiers may be cascaded with an over-all gain of man thousand, so that drifts of even a few millivolts at th input may be objectionable. After considerable exper mental and theoretical work on the matter, satisfacto amplifiers were developed. Since the main source of dri lies in the power supplies and the first stage of an ampl fier, it is necessary to use well-regulated power supplie and to choose carefully the tube for the first stage. Fil ment-type tubes with well-regulated filament curre were found superior to the well-known Miller circui However, since the simplest source of regulated voltag is the high-voltage power supply, a low-current tube desirable to avoid excessive power requirements. Son of the negative-gain servo amplifiers are shown in Fig. and the circuit diagrams for both types of amplifiers a shown in Fig. 8. In practice, the 1LN5 pentode with filament current of 0.040 ampere was found satisfacto for the negative-gain amplifier. A subminiature hearin aid tube, the CK522AX, is used for the positive-ga amplifier, since its filament current is only 0.020 an pere, while its drift and microphonism characteristics a very satisfactory. This tube cannot be used for th negative-gain amplifier because of its inability to pr vide a sufficiently large output voltage in a one-stag amplifier.

After about a two-hour warm-up period, the drift a referred to the grid of the first tube settles down to value that does not exceed 1 millivolt per hour. A typ cal drift record is shown in Fig. 9. The sharp pi

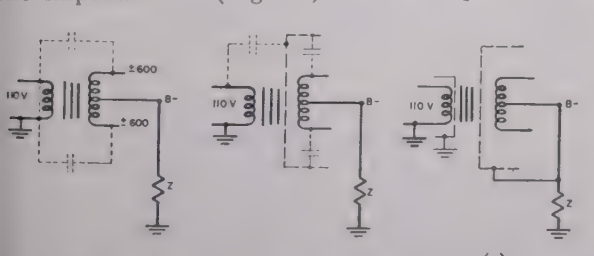


ig. 9—Typical drift characteristic of a negative-gain amplifier.

rked by X were due to supply-line switching surges. ce this drift stability exceeds the computer requirents, no attempt was made to obtain further improvent.

The other major problems in the design of the dc plifiers arise because the normal ground terminal —) of the amplifier cannot be connected to system ound unless isolators are used. It is conceivable that analogue circuit impedance to ground may be as high several thousand ohms, and that the voltage to ound may be as large as 30 or 50 volts. Two undesira- effects can arise. Unless the grid of the input tube the amplifier is very well shielded from true ground the $B-$ bus, the capacitance between grid and ound, in series with the normal high impedance from d to $B-$, results in a grid signal if a voltage exists ween $B-$ and true ground. It was considered desira- to have all amplifier chassis connected to true ound, so that shielding was required within the assis. Without resorting to complete electrostatic elds, the resulting transfer characteristics for the sitive-gain amplifier gave rise to an output voltage ual to approximately 0.1 per cent of the signal beeen $B-$ and true ground. This was considered satistory.

The other effect, a 60-cps ripple appearing on the $B-$ minal if an impedance is connected between $B-$ and e ground, is more difficult to eliminate. This ripple rmally arises from unbalanced capacitance between ends of the high-voltage winding and the primary nding, as shown in Fig. 10(a). With normal "balanced-nding" procedures, a ripple of about 0.1 volt appears the impedance Z (Fig. 10) is made equal to 1,000



(a) (b) (c)

ig. 10—Shielding requirements for power-supply transformers.

ohms. This corresponds to an unbalanced capacitance of a few hundred $\mu\mu$f. If a single shield, as shown in Fig. 10(b), is used, conditions are not appreciably improved. If, for example, the shield is connected to true ground, the unbalanced capacitances to the shield still exist. If the shield is connected to $B-$, then the capacitance between primary winding and shield still gives rise to an objectionable signal in the output. By careful choice of connections the ripple voltage may be reduced to 40 or 50 millivolts, but this is much larger than is permissible.

A very satisfactory solution results if two complete electrostatic shields are provided, as shown in Fig. (10(c). Because of construction difficulties, perfect shielding is not economically practicable. However, commercial manufacturers have provided transformers in which the unbalanced leakage capacitance is less than 0.05 $\mu\mu$f with an increase in cost ranging only from 50 to 100 per cent. This unbalanced capacitance gives rise to a ripple signal of about 0.1 millivolt per thousand ohms impedance.

The other characteristics of the amplifier are achieved by conventional means. The required low output impedance is achieved by use of a cathode-follower output and considerable negative feedback. The available output current is increased substantially above the value for a conventional cathode follower through use of a pentode as cathode impedance. The important characteristics of the amplifiers are listed in Table I.

TABLE I
PERFORMANCE DATA OF DC VOLTAGE AMPLIFIERS

	Negative-Gain Amplifier	*Positive-Gain Amplifier*
Gain	0.5 to 50 in. 1-db steps	1 to 100 continuously variable
Maximum voltage output	±125 volts peak	±125 volts peak
Maximum current output	±65 ma peak	±65 ma peak
Input impedance	400,000 ohms	500,000 ohms
Output impedance	25 ohms	0.9 ohms
Ripple level	1 mv rms	1 mv rms
Stability to line voltage fluctuations*	7 mv/10 volts	0.04 mv/10 volts
Drift characteristics*	1 mv/hour**	1 mv/hour**
60 cps ripple across 1,000 ohms between B- and true ground	1 mv	0.1 mv

* Voltage referred to grid of input tube.
** Typical values after 1-or 2-hour warm-up period.

DC Isolator and Preamplifier

Since isolators must, in general, operate with no connection to system ground, power transformers for such devices require the elaborate shielding discussed previously. Thus it is desirable for such isolators to operate without 60-cps ac power supplies, if convenient. On the other hand, the signal voltage available is frequently small, so that a preamplifier may be desirable to minimize the noise-to-signal ratio of the isolator.

Through the use of an amplitude-modulated carrier device, a very simple isolator results, with negligibly

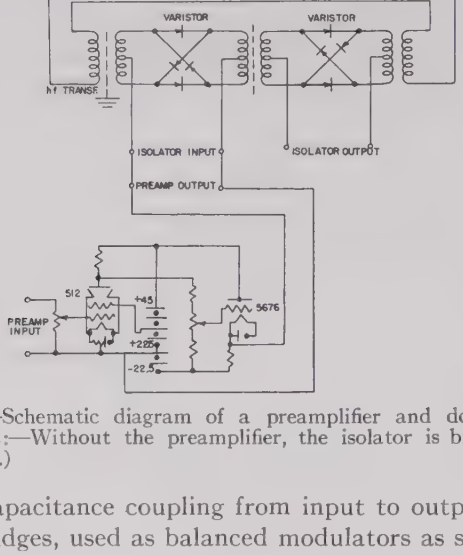


Fig. 11—Schematic diagram of a preamplifier and dc isolator. (NOTE:—Without the preamplifier, the isolator is bilateral in action.)

small capacitance coupling from input to output. Varistor bridges, used as balanced modulators as shown in Fig. 11, have given compact isolators with no 60-cps power supplies. The design of such a unit is not as straightforward as might be supposed, since the phase-shift requirement for the detected output carries over into a similar requirement regarding relative phase shifts of carrier and sidebands. However, through the use of nonresonant circuits, the phase-shift performance of the isolators can be made satisfactory.

A simple battery-operated amplifier is also shown in the circuit of Fig. 11. Both of these units can be mounted on a very small chassis. The isolator produces a useful signal of about 0.7 volt peak-to-peak and has satisfactory linearity and negligible phase shift in the range from zero frequency (dc) to 1,000 cps.

Generator of an Arbitrary Function of the Dependent Variable

A somewhat more difficult problem is to produce an arbitrary function of the dependent variable. As has been pointed out earlier, the dependent variable can be obtained as a voltage existing at some point in the analogue circuit. The problem is, then, to produce a voltage which is an arbitrary function of another voltage. Two devices have been developed for this purpose.

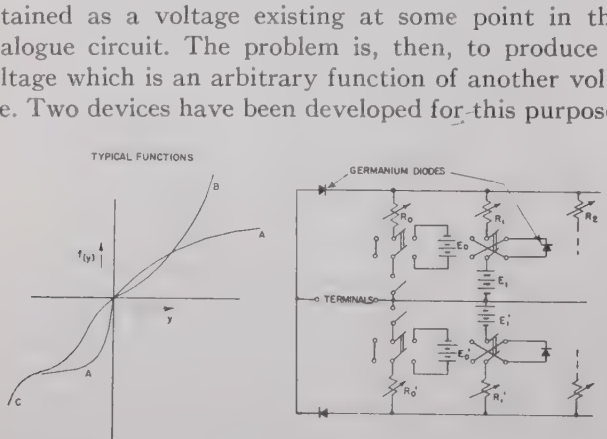


Fig. 12—Schematic diagram of a generator for the arbitrary function of a dependent variable—Type I.

One of the generators (Type I) utilizes crystal diodes and batteries in a manner shown in Fig. 12. The adjustable resistors are made relatively large compared to forward impedance of the crystals, so that the variati of the latter with current do not affect the over-all pedance appreciably. If the switches are all thrown the left in Fig. 12, each parallel branch will not cond current until the terminal voltage rises above the b tery voltage of that branch. The resulting current ver voltage characteristic is composed of a series of strai lines, with the slope of successive segments decreas as the voltage rises. Such a characteristic is shown curve A of Fig. 12.

If the switches are thrown to the right, a differ characteristic results. Each branch produces a circul ing current which passes through R_0. The battery vo age E_0 is provided to give zero terminal voltage when external current enters the terminals. As voltage is plied, the current which flows is determined by the fective resistance of all branches in parallel. As s as the terminal voltage rises above the lowest bra battery voltage, the current in that branch ceases to fl and the effective impedance of the device is raised. the voltage exceeds the highest battery voltage, impedance becomes equal to R_0 alone, the maxim value. The resulting current versus voltage charact istic is one of constantly increasing slope, as shown curve B. By combining branches of each type, a ch acteristic may be achieved whose slope is alternat increasing and decreasing, provided the slope alwa remains positive. Curve C in Fig. 12 is typical.

The generator of this type which is used with computer has 22 parallel branches, 11 for positive a 11 for negative voltages. A current-voltage charact istic composed of 22 straight-line segments can be tained. Such a characteristic can be made to appr imate many typical functions with surprisingly g accuracy. Simpler and more special-purpose versions this device have also been developed, in the form abrupt current and voltage limiters. These are shown Fig. 7 together with the Type I generator.

The other arbitrary-function generator (Type II based upon a principle developed elsewhere. It utili a cathode-ray tube and photocell. The quiescent p tion of the electron beam is set to one side of the t and the photocell output is amplified and applied to plates with a polarity which will drive the beam to other side. If an opaque template is placed on the s face of the tube, as shown in Fig. 13, the beam spot be driven below the edge of the template until the li reaching the phototube is just sufficient to produc signal which will deflect the beam to the edge of template.

The generator has several limitations which merit cussion. The finite spot size results in some distort for which compensation is difficult, since distortion pends on the gain of the amplifier, intensity of the sp the height of the pattern, location of the quiescent p tion, etc., all of which may vary from time to time. addition, the halo on the screen contributes an app

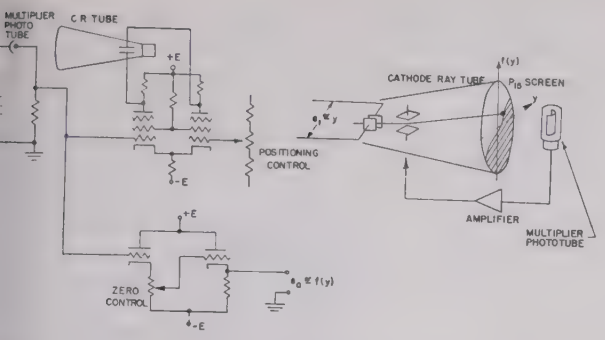


Fig. 13—Schematic diagram of a generator for the arbitrary forcing function of a dependent variable—Type II.

ble signal which may cause clipping of sharp high
ks of the template, since the halo signal can be
ked up on both sides of the peak with the spot com-
tely hidden. A third source of error lies in the paral-
resulting from a spot on one side of the cathode-ray-
e face and a pattern on the other side. Careful pro-
tioning of over-all height and width of the pattern,
ation of the quiescent position, and adjustment of
n minimize these difficulties, but a definite limitation
accuracy exists. Another limitation lies in the inabil-
of the beam to rise or fall abruptly. Thus, if the de-
dent variable passes rapidly through a region where
arbitrary function changes abruptly, the output may
behind the ideal curve. This effect can be minimized
a reduction in vertical scale of the template. Since
time lag is small compared to the period of highest
quency used on this computer, the effect on solutions
sually negligible.

The Type II function generator as developed for the
nputer is shown in Fig. 7. Its frequency response is
t defined in terms of the maximum speed of rise of the
hode-ray beam. Under typical operating conditions,
corresponds to 40 microseconds per inch of deflec-
n along the $f(y)$ axis. This figure was obtained with a
phosphor screen having a decay time of approx-
tely 20 microseconds. Screens of this type provide
iciently fast response characteristics for general use
the computer.

Multiplier

The multiplication of two arbitrary voltages can be accomplished in many ways. The method which has been adopted for the CIT computer utilizes a double-modulation and subsequent detection scheme (see the circuit of Fig. 14). A balanced modulator is used to produce sidebands through modulation of a carrier by one of the product terms. The resulting voltage is of the form $E_1 \cos(\omega_c t)$. This voltage is used as the carrier for a second modulator, which gives a sideband output $E_1E_2 \cos(\omega_c t)$. The carrier voltage is, of course, suppressed in both cases through the use of a balanced modulator. For convenience, a varistor bridge is used for the first, and electron tubes for the second modulator. Detection is accomplished with another variator bridge, minimizing the power-supply requirements of the unit. Isolation between inputs and output is automatically supplied by the high-frequency transformers. It will be noted, however, that the second input voltage is not isolated from system ground. If such isolation is required for both inputs, an additional isolator must be used. Since phase shift of the sidebands results in phase shift of the detected signal, care must be taken to minimize this effect. The first modulator, with untuned transformers, performs very well in this respect. Since tuned circuits are required in the second modulator, for several reasons, the phase shift here is minimized by overcoupling of the transformer in the plate circuit of the modulator.

In the computer, groups of the dc isolators and the multipliers are supplied by a common high-frequency generator, thereby permitting small, compact units.

Acknowledgments

Part of the development research on the electronic components of this computer was carried out under a Frederick Gardner Cottrell Research Grant of the Research Corporation. This grant covered research on the multipliers and arbitrary function of the dependent variable. The authors also wish to acknowledge the assistance of R. H. MacNeal and Werner Buchholz, who worked on this project under the Cottrell Research Grant.

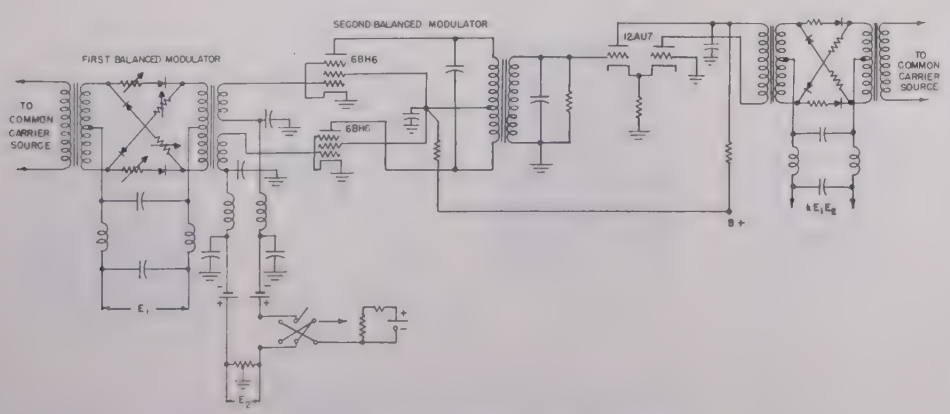


Fig. 14—Schematic diagram of a multiplier.

Contributors to Waves and Electrons Sectio

William S. Bachman (S'32–A'36–M'45) was born in Williamsport, Pa., in 1908. He was graduated from Cornell University in 1932, with the E.E. degree. Before he joined Columbia Records, Inc., in 1945, he was a member of the General Engineering Company's radio receiver engineering department, where he designed loudspeakers, developed AM and FM receivers and phonograph combinations, and perfected the GE Variable Reluctance phonograph pickup.

W. S. Bachman

❖

Richard H. Campbell, Jr. (S'47) was born in 1920, in Yonkers, N. Y. He received the B.S. degree in engineering sciences from Harvard University in 1942, and the M.Eng. degree from McGill University in in 1948.

Mr. Campbell was in the armed forces from 1942 to 1946, serving during the latter part of that period as a first lieutenant with the Army Security Agency. Since completing his course at McGill, he has been employed as development engineer by Minshall-Estey Organ, Inc., located in Brattleboro, Vt.

R. H. Campbell, Jr.

❖

Robert A. Chipman (A'41–SM'47) was born in Winnipeg, Manitoba, on April 28, 1912. He received the B.Sc. (E.E.) degree from the University of Manitoba in 1932, the M.Eng. (Comm. Eng.) degree from McGill University in 1933, and the Ph.D. (Physics) degree from Cambridge University in 1939.

From 1933 to 1935, he was demonstrator in physics at Cambridge University, and until 1937 he was demonstrator in physics at McGill University. From 1937 to 1938, he was a research associate at Cruft Laboratory, Harvard University. From 1938 to 1940, he was an assistant professor of physics at Acadia University, Wolfville, N. S. From 1940 to 1944, he was a research assistant and lecturer in physics at Queen's University, Kingston, Ont. He was an assistant professor of physics from 1944 to 1946 at Queen's University. During the period 1946 to 1947, he was an assistant professor of electrical engineering at McGill University, becoming an associate professor in 1947.

R. A. Chipman

Mr. Chipman is a member of the Acoustical Society of America, Sigma Xi, and the American Association of University Professors.

❖

Fred D. Clapp (S'43–A'44–M'45) was born on July 2, 1914, in Hanford, Calif. He became interested in radio at the age of 14 and obtained an amateur license shortly thereafter. In 1931, he decided to pursue radio and electronics as a life work.

There followed a succession of positions as police radio technician, broadcast station studio engineer, chief engineer and others. Upon the advent of war, Mr. Clapp entered the ESMWT training courses given at San Mateo Junior College and Stanford University in Calif., becoming an instructor during the first part of 1943. Upon termination of the program, he became associated with the University of California Radiation Laboratory, where he spent some time on electronic projects relating to the electromagnetic separation of uranium isotopes for the atomic bomb. A brief period near the end of the war was taken up as a radar field engineer for the Western Electric Company, New York, N. Y.

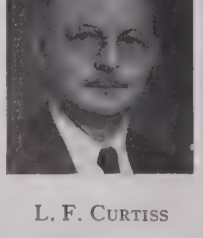
After the war, Mr. Clapp returned to the University of California, where he has since been employed as a research engineer in the antenna laboratory, operated under the department of electrical engineering.

Fred D. Clapp

❖

L. F. Curtiss was born in Adrian, Mich., on January 5, 1895. He received the A.B. degree in 1917, and the Ph.D. degree in 1922, both from Cornell University. During the period 1922–1924 he was a National Research Council Fellow at the Cavendish Laboratory in Cambridge, England. In 1924 he continued his Fellowship at the National Bureau of Standards, remaining until 1926, when he became a staff member as a physicist.

Dr. Curtiss was made chief of the Radioactivity Section in 1940, and holds that position at present. He has been chairman of the Committee on Nuclear Science of the National Research Council since 1946.



L. F. Curtiss

Peter C. Goldmark (A'36–M'38–F was born in Budapest, Hungary, in 1 He received the B.Sc. and Ph.D. deg at the Univer of Vienna. In 1 he came to United States, by the end of 1 had been made c television enginee the Columbia Bro casting System. present, Dr. G mark is director engineering resea and development CBS.

Peter C. Goldmark

Dr. Goldmark was the recipient of IRE Morris Liebmann Memorial Prize 1945 for his work on color television. H also a Fellow of the Society of Motion ture Engineers.

❖

For a biography and photograph Eugene F. Grant see page 877, of July, 1948, issue of the Proceedings of t I.R.E.

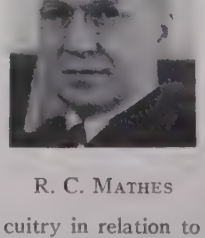
❖

B. N. Locanthi (S'38–A'48) was born White Plains, N. Y., on September 10, 19 He received the B.S. degree from the C fornia Institute Technology in 19 He was a resea associate during 1 designing electro equipment for t metering data fr bombs in flight. present, he is a velopment engin at the analysis l oratory of the C fornia Institute Technology.

B. N. Locanthi

❖

R. C. Mathes (M'42–SM'43) was b on April 19, 1891, at New Ulm, Minn. was graduated from the University of M nesota with the grees of B.Sc. in 19 and E.E. in 1913. joined the resea department of Western Elec Company in 19 and continued in same department Bell Telephone La oratories from 19 to date. Mr. Mat has been principa concerned with cuitry in relation to communications pr lems, including early application of vacu tubes to telephone repeaters and ocean ca



R. C. Mathes

raphy; voice-operated circuits for echo ressors, radio terminals, automatic me control, and compandors; picture smission and television; privacy and cy systems; narrow-band transmission ems; negative impedance circuits; and application of electronic methods to ching and signaling circuits.

Mr. Mathes is a member of Tau Beta nd the American Physical Society; and llow of the American Association for Advancement of Science, the Acoustical ty of America, and the American Inte of Electrical Engineers.

❖

Gilbert D. McCann was born in Glen- Calif., on January 12, 1912. He received B.S. degree in 1934, the M.S. degree in 1935, and the Ph.D. degree in electrical engineering in 1939 from the California Institute of Technology. From 1938 to 1941, he was a central station engineer at Westinghouse Electric Corp. He became a transmission engineer in charge of lightning research from 1941 1945, and then became the consulting smission engineer in charge of the ning research and analytical laboratory e Westinghouse Electric Corp.

D. McCann

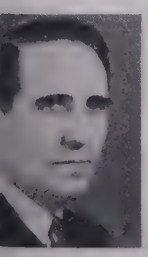
From 1946 to 1947, Dr. McCann was an ciate professor of electrical engineering ne California Institute of Technology, ming a professor in 1947. He was a tinghouse lecturer at the University of burgh from 1941 to 1946.

Dr. McCann is a member of the Ameri- Institution of Electrical Engineers and American Society of Mechanical Eners. He is a member of several AIEE nittees, including basic sciences and e-scale computing devices. He is a memof the IRE committee on electronic puters.

Dr. McCann is the author of a book and erous technical papers, chiefly on ning, power transmission, and analog puters.

❖

D. A. McLean was born in Golden, Colo., uly 15, 1905. Since graduation from the ersity of Colorado in 1929 with the degree of B.S. in chemical engineering, he has been a member of the technical staff of the Bell Telephone Laboratories. He has been primarily concerned with the chemical aspects of electrical insulation with particular reference to capacitor dielectrics, and has published a number of papers in this field. Among the patents he has been granted is one covering the now widely used method of stabilizing chlorinated capacitor impregnating compounds by the addition of small quantities of anthraquinone or similar materials.

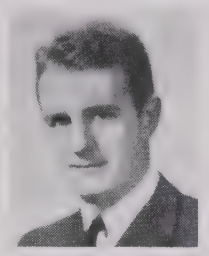


. A. McLean

Mr. McLean is a member of the American Chemical Society and a fellow of the American Institute of Chemists. Among various honorary and professional societies, he holds membership in Tau Beta Pi and Sigma Xi.

❖

William S. McLean was born in Baltimore, Md., on March 19, 1919. He was graduated from The Johns Hopkins University in 1941 with a degree in electrical engineering.



William S. McLean

During the period 1941–1942, he was employed by the General Electric Co., Schenectady, N. Y. on the Student Test Course. During this time, he worked in the Pittsfield, Schenectady, and Philadelphia plants of General Electric, testing various types of electric power equipment. In 1942 he went on active duty in the United States Navy as a radar maintenance officer specializing in fire-control radar, attended several of the Navy radar and fire-control schools, and later served two years in the Southwest Pacific doing maintenance work.

Upon his discharge from the service in 1945, Mr. McLean was employed by National Union Radio Corporation in Orange, N. J., where he worked on the design and development of various types of secondary-emission tubes. In 1948 he was employed by the radio division of the Western Electric Co. in their field engineering force, and is now engaged in radar installation and servicing work.

❖

Carl F. Miller (A'28–M'47–SM'48) was born in Germany on January 17, 1901. He was graduated from State College, Stuttgart, Germany, in 1926, with a degree in mechanical engineering. In 1927 he emigrated to the United States, and he became a citizen in 1933.

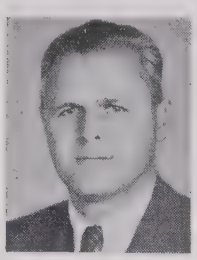
Carl F. Miller

From 1927 to 1929, he was employed by the Federal Telephone and Telegraph Co., Buffalo, N. Y., as an engineer for the development of superheterodyne receivers. He joined the Sylvania Electric Products Co., Emporium, Pa., in 1929, and developed the first 6.3-volt line of tubes for automobile receivers; later he invented the lock-in tube and supervised the early production stage. In 1940 he became associated with the Corning Glass Works, Corning, N. Y., as design engineer for their multiform division.

National Union Radio Corp., Orange, N. J. appointed Mr. Miller as chief engineer of their Power Tube Plant in 1943, and since then he has been engaged in the development of electron tubes of various types.

❖

H. A. Sauer was born on Staten Island, N. Y., on May 3, 1908. He attended Cooper Union and New York University, from which institutions he received the B.S. degree in chemical engineering in 1930, and the M.S. degree in physics in 1939, respectively.



H. A. Sauer

As a member of the technical staff of the Bell Telephone Laboratories, Mr. Sauer has been active principally in the field of dielectrics, and in the development of test equipment and techniques. He is a member of Sigma Pi Sigma and the American Physical Society.

❖

René Snepvangers was born in Antwerp, Belgium, in 1900. He attended various Belgian Universities until 1922. From 1928 to 1934 he was employed as an engineer with the Bell Telephone Manufacturing Company in Antwerp; then he became a consulting engineer for the Belgian Branch of the United Electrical Works, "Tungsram," of Budapest. In 1939, he came to the United States, joining the RCA Research Laboratories as a specialist in phonograph development. In November, 1944, Mr. Snepvangers became associated with CBS Recording Laboratories, and is now the senior recording research engineer.

René Snepvangers

❖

Charles H. Wilts was born in Los Angeles, Calif., on January 30, 1920. He received the B.S. degree in 1940 and the Ph.D. degree in 1948 in electrical engineering from the California Institute of Technology. From 1945 to 1947, he was a National Research Council Fellow. He is the author of three papers on application of analogue computers to engineering problems. Presently, he is an assistant professor of applied mechanics at the California Institute of Technology.

C. H. Wilts

Abstracts and References

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ACOUSTICS AND AUDIO FREQUENCIES

534.21 1839

On the Sound Field Produced by a Point Source with Uniform Rectilinear Ultrasonic Velocity in a Perfect Fluid—P. Liénard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 910–912; March 14, 1949.

534.22:534.321.9 1840

Experimental Determination of Velocity of Sound in Superheated Steam by Ultrasonics—J. Woodburn. (*Trans. Amer. Soc. Mech. Eng.*, vol. 71, pp. 65–70; Discussion, pp. 70–72. January, 1949.) The frequency of a quartz crystal source and the length of standing waves generated in a cylinder containing the steam were measured. Results agree closely with data calculated from the steam tables of Keenan and Keyes.

534.232 1841

Coupled Mechano-Acoustic Oscillators and Resonators—F. A. Fischer. (*Frequenz*, vol. 2, pp. 232–238; September, 1948.) Discussion, by means of electrical analogies, of the type of oscillator termed "Tonpilz" [literally "sonic mushroom"] by Hahnemann and Hecht (see *Phys. Z.*, vol. 21, pp. 187–192, 1920; and vol. 22, pp. 353–360; 1921) and consisting simply of two masses connected by a spring and a damping device. The coupling of two such systems is exemplified in the well-known Fessenden oscillator. Reciprocal arrangements, in which two resonators are connected by a neck, are also considered.

534.321.9 1842

Experimental Ultrasonics: Parts 1 & 2—S. Y. White. (*Audio Eng.*, vol. 33, pp. 20–23, 45 and 24–25, 41; March and April, 1949.) Part 1: Design, construction, and operation of the Hartmann whistle. Part 2: Discussion of small pickups such as those used for probes, with due allowance for the wide frequency range normally required.

534.321.9:537.228.2 1843

Ultrasonics and Electrostriction—H. Falkenhagen. (*Z. Angew. Phys.*, vol. 1, pp. 304–306; March, 1949.) Discussion of various possible methods of using electrostriction in ultrasonic transmitters and receivers.

534.41:534.78 1844

A Photoelectric Type of Acoustic Spectrograph Using Sound Film—D. Brown, C. F. Coleman, and J. W. Lyttleton. (*Proc. Phys. Soc.*, vol. 62, pp. 149–162; March 1, 1949.) An optical method in which a continuously moving sound film is analyzed by means of a frequency-scanning disk; the resulting oscillograph pattern is photographed on another moving film. Typical analyses, mainly speech patterns, are reproduced and discussed. Advantages include the ability to record the amplitude and phase of each component, and to change the analyzing convention in any desired manner. Resolving power, spectral line profiles, and amplifier bandwidth requirements are also considered.

534.43:621.395.813 1845

Cost vs. Quality in A.F. Circuits—J. M. Van Beuren. (*FM-TV*, vol. 9, pp. 31–32, 34; February, 1949.) "Carefully-chosen components and well-designed circuits can give a big improvement [in quality] at very small cost." Three high-quality af amplifiers are discussed, with diagrams and component details.

534.76:621.395.625.6 1846

Experiment in Stereophonic Sound—L. D. Grignon. (*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 280–292; March, 1949. Discussion, p. 292.) S.M.P.E. 1948 Convention paper. Discussion of the special microphone technique required for recording for reproduction in a cinema. Essentially the same paper is reproduced in *FM and Telev.*, vol. 9, pp. 28–30; April, 1949.

534.78:621.396.619.13/.14 1847

Ratio of Frequency Swing to Phase Swing in Phase- and Frequency-Modulation Systems Transmitting Speech—Gannett and Young. (*See* 2048.)

534.83 1848

Noise and its Measurements—H. G. M. Spratt. (*Elec. Rev.* (London), vol. 144, pp. 565–567; April 8, 1949.) A short account of the principal features and functions of the sound-level meter Type 1400, af analyzer Type 1401, and vibration meter Type 1402, made by the Dawe Instrument Co.

534.844.4

A Comparison of the Acoustics of the F harmonic Hall, Liverpool, and St. Andre Grand Hall, Glasgow—T. Somerville. (*Quart.*, vol. 4, pp. 41–54; April, 1949.) acoustic properties of the halls are expresse terms of reverberation time, and various m ods are discussed for measuring this quant which depends on the frequency of meas ment and the alignment of the directi microphone used. The results are correla with the constructional details of the halls, with subjective observations.

534.845

On the Theory of the Reflection of So by Porous Media—J. Korringa, R. Kronig, A. Smit. (*Physica, 's Grav.*, vol. 11, pp. 209– December, 1945. In English.) An extensio recent investigations by Zwikker and his laborators (3321 to 3324 of 1948) to a med consisting of a large number of identical r spheres, with infinite heat capacity and ther conductivity, arranged in a cubic lattice.

534.86:621.397.5

Audio Technique in Television Broadc ing—R. H. Tanner. (*Audio Eng.*, vol. 33, 9–13, 44; March, 1949.)

621.395.61/.62

Considerations on Electroacoustic Tr ducers and Some Conclusions, with Prac Examples—W. Bürck. (*Funk und Ton*, vo pp. 187–201; April, 1949.) General theor electrodynamic, electromotive, and elec static devices, with particular reference efficiency. The conclusions are applied to cussion of the dynamic loudspeaker and dynamic microphone.

621.395.61

An Omnidirectional Microphone—J. Hilliard. (*Audio Eng.*, vol. 33, pp. 20 April, 1949.) The microphone is 0.6 inc diameter and 0.4 inch thick; it weighs less ¼ oz. It is based on electrostatic rather electromotive principles, and is little affe by blast, iron filing damage, or magnetic in tion. In sound reinforcement applications, more gain is obtainable than with other directional microphones.

621.395.61

Single-Element Unidirectional Micropl—H. F. Olson and J. Preston. (*Jour. Soc. Pic. Eng.*, vol. 52, pp. 293–302; March, 19 Developed for use in sound-film recording, the following characteristics: single-ri type; the back of the ribbon is coupled damped folded pipe and an acoustical im

in the form of an aperture; improved ioid directional pattern; increased output; ced weight; reduced wind-noise response.

395.623.7 1855
Modern Loudspeaker Technique—M. ant. (*Radio Tech. Dig.* (Franç), vol. 3, 33–99; April, 1949. Bibliography, pp. 99–) A short discussion of the fundamental ry of electrodynamic loudspeakers, with lated data for the principal permanent-net alloys and lists giving some details of anent-magnet and electromagnetic loud-kers manufactured in France. Modern s of American and British loudspeakers are wed briefly.

395.625.2:621.396.712 1856
Disc Recording for Broadcast Stations—W. ahoney. (*Audio Eng.*, vol. 33, pp. 9–13, 46; , 1949.) Discussion of the system designed ew studios at WSAI, Cincinnati, and of the ical problems involved in designing a lex system of recording equalizers.

ANTENNAS AND TRANSMISSION LINES

15.212:621.397.5 1857
The London-Birmingham Television Cable: 2—Cable Design, Construction and Test lts—H. Stanesby and W. K. Weston. *Elec. Eng. Jour.*, vol. 42, part 1, pp. 33–April, 1949.) Details of the experimental on the design and manufacture of the , the production of factory lengths, and pointing technique used for completing re-r sections. Test results on typical sections lso discussed. Part 1: 1279 of June.

92.26† 1858
eflection of H_1 Waves at a Sudden ge of the Cross-Section of a Rectangular eguide—K. Fränz. (*Frequenz*, vol. 2, pp. 231; September, 1948.) A formula for the tion coefficient is derived which depends e waveguide dimensions and is analogous at for reflection at the junction of two lel Lecher-wire systems of the same lth but with their ends at different levels. e reflection point, a local capacitive field .

96.67 1859
he General Problem of Antenna Radiation he Fundamental Integral Equation, with cation to an Antenna of Revolution: 1 & 2—G. E. Albert and J. L. Synge; J. nge. (*Quart. Appl. Math.*, vol. 6, pp. 117–July, 1948.) Part 1: Radiation in a finite y is discussed, and an integral equation is ned which has the same form as the basic ral equation for an antenna. The latter ion reduces to a simple explicit form for an na with axial symmetry. The application e results to actual radiating systems is dis-d. Part 2: The importance of the gap is asized; for a fuller investigation see 24 of (Infeld). Thin antennas are considered; current is shown to be approximately oidal outside the gap. A shape term is in-d for the case when the gap is not at the r, and is calculated explicitly for certain e shapes. A method of successive approxi-ons is also described for dealing with any na, thin or thick.

96.67 1860
he Helical Antenna—J. D. Kraus. (Proc. ., vol. 37, pp. 263–272; March, 1949.) The is considered as a fundamental form of na and the variation of radiation pattern ype of polarization with change in helix nsions are discussed. For the axial mode, adiation pattern and terminal impedance aintained over a wide frequency range.

621.396.67 1861
A V.H.F. Helical Beam Antenna—H. E. Taylor and D. Fowler. (*CQ*, vol. 5, pp. 13–16; April, 1949.) Design details for an antenna for the 144- to 148-Mc range, with circular polarization and gain of 11.5 to 13 db.

621.396.67 1862
Optimum Design of a Cylindroparabolic Reflector—G. Klages. (*Frequenz*, vol. 2, pp. 151–154; June, 1948.) The Kirchhoff-Huyghens principle is used to calculate the zones on the surface of the reflector which reduce, by interference, the radiation along the axis. The formulas obtained enable reflectors to be designed with optimum gain.

621.396.67:538.56:535.13 1863
The Radiation and Transmission Properties of a Pair of Semi-Infinite Parallel Plates: Part 1—Heins. (*See* 1920.)

621.396.67:621.396.11 1864
Ground Absorption with Elevated Vertical and Horizontal Dipoles—R. E. Burgess. (*Wireless Eng.*, vol. 26, pp. 133–139; April, 1949.) The fraction of the power absorbed in the ground is calculated for dipole heights which are large compared with the wavelength. The dipole is assumed to have a sine-law polar diagram and the energy flow into the ground is evaluated using the Fresnel reflection coefficients which depend upon the wave polarization. The absorption factors are determined for limiting values of ground permittivity. The ground absorption for a vertical dipole is a very slowly varying function of the electrical properties of the ground and is very nearly 0.4 for most practical conditions. It is appreciably greater than that of a horizontal dipole on account of the Brewster phenomenon. The results obtained are compared with those derived earlier by Strutt, Niessen, and Sommerfeld and Renner.

The noise in an antenna system due to thermal radiation from the ground is evaluated directly using thermodynamical principles and is shown to be consistent with the value derived by application of the reciprocity principle.

Brief consideration is given to the ground absorption for antenna arrays.

621.396.671 1865
Driving Point Impedance of a Vertical Cylindrical Radiator and Concentric Ring of Subsidiary Radiators over Perfectly Conducting Earth: Parts 1 & 2—H. Cafferata. (*Marconi Rev.*, vol. 12, pp. 12–20 and 57–67; January to March and April to June, 1949.) Hallén's integral equation method is applied, with particular reference to an approximate solution for the case of 3 small subsidiary radiators symmetrically arranged round the main antenna and with base connections to it, the radius of the ring being small compared to λ. See also 854 of 1946 (Harrison).

621.396.671 1866
Measurement of Gain of Electromagnetic Horns—A. S. Dunbar and M. D. Adcock. (*Jour. Appl. Phys.*, vol. 20, pp. 226–227; February, 1949. Criticism of 1600 of July (Watson and McKinney), and discussion of an alternative method involving measurement only of the ratio of received to transmitted power.

621.396.671 1867
Mutual Impedance of Parallel Aerials—G. Barzilai. (*Wireless Eng.*, vol. 26, p. 73; February, 1949.) Corrections to 648 of April.

621.396.677 1868
Metallic Delay Lenses—S. S. D. Jones and J. Brown. (*Nature* (London), vol. 163, pp. 324–325; February 26, 1949.) Kock's formula (2176 of 1948) for the refractive index n is found experimentally to give a value 18 per cent too low. The new formula

$$n^2 = 1 + (2d/\pi a)\log_e \operatorname{cosec}(\pi b/2d),$$

where a, b and d are dimensions of the lens structure, is based on the assumption that the refracting medium behaves like a transmission line. The formula is valid for the wavelength range within which n is constant. Observed values of n agree with those predicted by this formula within the limits of experimental accuracy.

621.396.677 1869
Directional Antenna for the 152–162-Mc Communications Band—J. S. Brown and V. J. Moffatt. (*Communications*, vol. 29, pp. 14–16, 35; March, 1949.) A corner-reflector type with vertical polarization, in which two sets of parallel rods, spaced 0.1 λ apart in two vertical planes, are used instead of metal sheets. It has a unidirectional radiation pattern, a gain of at least 6 db over a λ/2 dipole and a voltage SWR < 1.5 over the whole frequency band. The feeder cable can have any impedance between 50 and 75Ω.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3:621.314.6 1870
Voltage Surges Produced in Rectifiers by Starting without Load—T. Douma. (*Communication News*, vol. 9, pp. 121–125; August, 1948.) If a high-voltage rectifier with low damping is started without load, voltage surges up to four times the peak transformer voltage can occur across the tubes, and the voltage on the smoothing capacitors can be twice the peak alternating voltage of the transformer. These surge voltages are at least double those encountered under normal operating conditions. They can be avoided by applying damping to the input choke and providing the smoothing capacitors with a suitable series resistance which can be short-circuited after the rectifier has been switched on.

621.3.09:621.397.5 1871
Attentuation and Phase Distortion and Their Effect on Television Signals—G. Fuchs and V. Baranov. (*Câbles and Trans.* (Paris), vol. 3, pp. 194–207; April, 1949.) An examination of the relations between small attenuation and phase distortions and the deformation of the response curve of a television receiver. In one method, the distortion versus frequency characteristic is treated as a whole, while in a second method this characteristic is split up into a certain number of elementary sections. Practical limits for these distortions are calculated and the theoretical results are applied to 450-line and 1,100-line television systems.

621.314.2.012.8 1872
The Transformer and its Equivalent Representation—P. G. Violet. (*Frequenz*, vol. 3, pp. 1–12; January, 1949.) A general treatment based on quadripole theory. Matrices for the ordinary and the ideal transformer are given and the various equivalent schemes are explained and compared. The application of circuit equivalents for measurement purposes and for filter calculations is considered, in particular for filters whose branches consist of oscillatory circuits. The equivalents for transformers with two secondary windings are given and comparison is made with the normal single-secondary transformer.

621.314.3† 1873
The Basic Considerations in the Design of Push-Pull Magnetic Amplifiers—M. A. Rozenblat. (*Avtomatika i Telemekhanika*, vol. 10, pp. 32–50; January and February, 1949. In Russian.) Discussion of amplifiers of the differential bridge, and transformer types (Figs. 1, 2, and 3 respectively) with loads taking ac. Their operation is discussed and the conditions for obtaining maximum values of both power output and

amplification factor are established. Methods are indicated for determining the optimum operating conditions for a given core material. An example of the design procedure is given, together with experimental results for the amplifier so designed.

621.316.86 1874

Cracked-Carbon Resistors—R. W. Wilton. (*FM-TV*, vol. 9, p. 29; February, 1949.) Brief discussion of the characteristics of resistors made by the Welwyn Electrical Laboratories, by the pyrolitic process of depositing carbon on porcelain. Nonlinearity hardly exists. The resistance value is controlled by the cracking temperature, the hydrocarbon content of the gas mixture, and the time of exposure of the rod in the cracking zone. Such resistors are cheaper than wire-wound resistors for ratings up to 2 to 4 watts, and inductive effects are almost absent from high-value resistances, for which helical grooves may be used. Resistance values are accurate within ± 1 per cent, voltage and temperature coefficients are low, and long-term stability is high. High-gain amplifiers using these resistors have little high-level hiss.

621.316.86 1875

Results of an Investigation of the Volt-Ampere Characteristics of Some Nonlinear Resistors—M. A. Topchibashev. (*Avtomatika i Telemekhanika*, vol. 10, pp. 13–24; January and February, 1949. In Russian.) The division of nonlinear resistors into two main groups (varistors and thermistors) is considered; their properties can be best represented by their volt-ampere characteristics. The equation describing these is of the form $I = AV^{\alpha}$ where A is called the coefficient of nonlinearity and α the index of nonlinearity. Analytical and graphical methods for determining A and α are discussed. Results of an experimental investigation of a number of these resistors are summarized in two comprehensive tables showing A, α, and the maximum errors for given voltage ranges.

621.318.42 1876

Axial Field and Supplementary Losses of Inductance Coils without Iron—J. Rezelman. (*Rev. Gén. Élec.*, vol. 58, pp. 154–162; April, 1949.) Formulas and curves are given for circular coils of rectangular winding cross section with experimental results.

621.318.572 1877

A New Pulse-Amplitude Discriminator Circuit—E. H. Cooke-Yarborough. (*Jour. Sci. Instr.*, vol. 26, pp. 96–97; March, 1949.) The circuit described discriminates against pulses smaller than those to be counted. The dead time is a known constant, so that counting losses can be accurately determined without complete knowledge of the resolving time of other equipment in the counting chain. These counting losses are calculated and the condition for the counting rate to be independent of the input pulse width is deduced.

621.318.572:621.396.1 1878

An Electronic Switch for Diversity Reception—H. V. Griffiths and R. W. Bayliff. (*BBC Quart.*, vol. 4, pp. 57–64; April, 1949.) The dual-diversity switch has a discriminator which monitors the signals received on two receivers and controls a gating circuit which blocks the output from the receiver carrying the weaker signal and passes that from the receiver carrying the stronger signal. Circuit details are given of this switch and of a similar device for operation with three receivers (triple-diversity).

621.319:679.5 1879

Design and Performance Characteristics of "Electrets"—E. D. Padgett. (*Tele-Tech*, vol. 8, pp. 36–39; 63; March, 1949.) See also 1307 of June.

621.319.4 1880

Capacitors with Cylinders of Different Expansion Coefficients for Compensation of the Dependence of Capacitance on Temperature—A. Rohrmann. (*Funk und Ton*, vol. 3, pp. 230–233; April, 1949.) Design formulas are given, with numerical examples and tabulated temperature coefficients of capacitance for capacitors with outer cylinder of brass (37 per cent Zn) and inner cylinder either of copper or of brass with Zn content ranging from 28.1 to 37 per cent. The coefficient is zero for a capacitor with Cu inner cylinder and radii ratio of 1.164, and also for one with brass inner cylinder (31.1 per cent Zn) and radii ratio of 1.02.

621.319.4:621.315.59 1881

On the Use of Semiconducting Liquids for Impregnating Paper Capacitors—Renne. (*See* 1969.)

621.392 1882

General Impedance-Function Theory—P. I. Richards. (*Quart. Appl. Math.*, vol. 6, pp. 21–29; April, 1948.) Discussion of methods of extending lumped-constant impedance-function theory to distributed-constant circuits.

621.392:621.396.619.13 1883

A Simple Method of Calculation for Electrical Circuits Carrying Frequency-Modulated Voltages—J. W. Alexander. (*Communication News*, vol. 9, pp. 33–38; December, 1947.) An approximate simplification of Vellat's method (79 of 1942), accurate in general within 10 per cent for the final result. The important parameters governing the accuracy are the "swing-factor," or ratio of the maximum frequency deviation to the maximum modulation frequency, and the bandwidth factor, or ratio of the bandwidth to the double swing. Calculations are made as for a circuit in which the voltage varies sinusoidally at frequency f, and f is replaced by the instantaneous frequency in the final result. The distortion for a tuned circuit with and without limiter, and for a bandpass filter, is calculated by this and by other methods and the results are compared.

621.392.43 1884

A Balanced-to-Unbalanced Matching Unit for High Frequencies—J. W. Whitehead. (*Jour. Sci. Instr.*, vol. 26, pp. 71–73; March, 1949.) A circuit for matching an unbalanced circuit to a balanced resistive circuit is investigated theoretically and practically. Component values for a particular case are derived. Experimental units on the same principles have been built with certain inductances and capacitances variable; they have been used successfully with a 3.5-kw transmitter throughout the frequency range 2.5 to 20 Mc. These units can easily be constructed, and adjustments for large changes of frequency can quickly be made since coil changing is unnecessary.

621.392.43 1885

Impedance Transformation by Means of Cables—W. Burkhardtsmaier. (*Funk und Ton*, vol. 3, pp. 151–167 and 202–213; March and April, 1949.) Discussion of the use of sections of cable, with characteristic impedance variable along their length, as impedance transformers. Two types are considered; in the first, the impedance varies stepwise from one end to the other, while in the second, the impedance variation is uniform. Such transformers can be used with advantage not only for long and medium waves, but also for short waves if suitably designed. Design formulas are derived and practical types of equipment for concentric cables and twin conductors are described. In many cases, better results can be obtained by the addition of compensation units at each end of the transformer section. The design of such units is also considered.

621.392.5

On the Synthesis of the Most Ge Passive Quadripoles—R. Leroy. (*Câbles Trans.* (Paris), vol. 3, pp. 141–158; A 1949.) The matrices of the impedances o mittances of passive quadripoles are "pos real" matrices. The construction of quadri from such matrices is straightforward their elements comprise only two types o pedance, but this is not the case when matrices include resistances, inductances, also capacitances. The possibility of obtai passive quadripoles corresponding to matrices was shown by Gewertz (*Jour. Phys.*, vol. 12, pp. 1–254; 1932–1933) w paper appears to-day to be largely ignore technicians. Gewertz studies a particular of matrices in which the determinant o even parts of the elements is zero and for w he gives a method of reduction. He also s that the representation of every "positive matrix can be referred to a matrix of this

The present paper, by more systemati plication of the theory of functions and of tive real" matrices, simplifies the resul Gewertz and gives physical reasons for ce properties, sometimes established by fo calculus, which may not appear altog satisfactory. Gewertz' method is also ap to the realization of effective impedances also 2539 of 1940, 1627 and 3017 of (Cauer) and 1887 below.

621.392.5

Reactive Ladder Networks—J. (*Câbles and Trans.* (Paris), vol. 3, pp. 159 April, 1949.) A theoretical analysis based o properties of the input impedance of la networks terminated by an ohmic resist This impedance only defines a network w limited number of cells, which are all-pass pass cells which cannot be realized by la construction are essentially of the lattice

Systematic use is made of shunt freque for which the shunt arms of the ladder net are short-circuited, cutting out the ter ohmic resistance; the input impedance the sembles a reactance and the shunt freque can, therefore, be determined *a priori* by ing the frequencies for which the input im ance has the properties of a reactance. shunt frequencies are roots of the equ whose first member is obtained by equati zero the even part of the input imped Comment on certain parts of the paper is by R. Leroy. See also 2190 of 1948 and above (Leroy).

621.392.52

Introduction to the Operating-Para Theory of Filter Calculation—C. Wisspei (*Frequenz*, vol. 2, pp. 131–140, 154–163, 199, and 210–214; May to August, 1948.) principal results of the mathematical the filters are presented in as simple a form as ble to make them readily available to the tical engineer. Quadripole theory is first sidered and in this the usual complex quan are denoted by German characters, Roman type is used in the treatment o operating-parameter theory. The wave-pa eter theory of quadripoles includes discu of the properties of various types of filte practice, it is seldom possible to termin filter with its characteristic impedance a consequence, reflections occur which affe attenuation both within and outside the band. The operating-parameter theory account of such reflections. The relatio tween over-all, current, and voltage att tion is discussed and also echo attenuation determination of the no-load resistance given attenuation is explained in detail the aid of numerous diagrams. Applicati the theory is illustrated by numerical ca tions for many different types of filter. R

a particular band-pass filter are in fairly l agreement with measurements.

392.52 1889
Variable Bandwidth Crystal Filters—B. del. (*Radiotronics*, pp. 78–87; September October, 1948.) The behavior of the ivalent circuit of a quartz crystal is dis-ed. For best over-all selectivity, the shunt acitance in this circuit should be neutralized practical arrangements for achieving this described. To vary the bandwidth, a tuned uit should be used as the load for the filter; dynamic impedance of this circuit can be red by the insertion of series or parallel stors. The stage gain is calculated, with due wance for bandwidth change. A design mple is discussed in detail.

392.52 1890
The Input Impedance of Some Low Pass ers with Resistance Terminations, with erence to Class "B" Modulator Applica-s—H. R. Cantelo. (*Marconi Rev.*, vol. 12, 41–56; April to June, 1949.) The properties certain *m*-derived filters terminated by a of constant resistance are examined at uencies up to 1.5 times the cutoff fre-ncy. With a preferred configuration of filter, secondary leakage inductance of the modu-on transformer can be included as part of filter system. Formulas and curves are n for 3 types of filter. The attenuation and se shift between input and output of the r system are also calculated.

392.52:621.395.44 1891
Telephony Filters and the Effect of Ter-ation and Losses on Their Characteristics: ts 1 & 2—J. F. Schouten. (*Communication ws*, vol. 9, pp. 61–69 and 97–106; April and ust, 1948.) If we assume that all the ele-nts of a reactive network have the same loss-or, the problem of calculating the effect of se losses on the network can be reduced to of determining a function of a complex ument when the function is known for an ginary argument. Three methods of solution discussed: (a) the method of the perpendicu-derivative, based on the differential expres-s of Cauchy and Riemann, (b) representa-, based on Laplace's differential equations, he attenuation and phase of a filter by ns of soap-film models, and (c) an inter-ation, based on Cauchy's integral, anal-us to a blurred image in optics.
Within certain limits, a relation exists be-en the attenuation and phase in a network, function of frequency; this relation is ul-ately based on the causal relation between input and output voltage. The soap-film hod is applied to several common types of ple filter which can be considered as proto-es of more complex filters.

392.52:621.396.621:621.396.5 1892
Application of Crystal Filters in [Telegra-] Receivers—F. Maarleveld. (*Tijdschr. Radiogenoot.*, vol. 14, pp. 41–55; March, 9. Discussion, p. 56. In Dutch with English mary.) The filters are designed for an f 100 kc and bandwidths of 50, 250, 500, 1,500 cps respectively. Design formulas tabulated. See also 2041 below.

395.667:621.397.6 1893
Phase and Amplitude Equalizer for Televi-n Use—E. D. Goodale and R. C. Kennedy. *A Rev.*, vol. 10, pp. 35–42; March, 1949.) scribes an equalizer of variable frequency onse. Two amplifiers having resonant-cir-loads of variable damping and resonance uency are used for amplitude equalization. compensate for the phase distortion in the plied signal and for that produced in the plitude equalizer, amplifiers with variable se shift and constant gain are used.

621.396.611.3 1894
Oscillator Power Variation and Frequency Pull-In—L. S. Schwartz. (*Tele-Tech*, vol. 8, pp. 30–32, 57; January, 1949.) An analysis of the effect of coupling variation for a mismatched feeder and an oscillator. Curves show how the oscillator power varies with coupling coefficient and SWR. Results of measurements using a lighthouse triode and cavity resonator are also included.

621.396.615 1895
Frequency Variation of a Crystal-Bridge Oscillator—W. Herzog. (*Arch. Elek.* (Übertragüng), vol. 2, pp. 357–361; December, 1948.) A simple bridge arrangement in the grid circuit of a tube oscillator enables a considerable variation of the crystal resonance frequency to be achieved. An auxiliary parallel circuit may be used to suppress one of the two possible oscillation frequencies.

621.396.615 1896
Variable Frequency R-C Oscillators—F. Butler. (*Electronic Eng.*, (London), vol. 21, pp. 140–142; April, 1949.) Discussion of phase-shift types having a frequency-selective network associated with a suitable amplifier. A practical oscillator circuit for the frequency range 250 to 5,000 cps is described in detail.

621.396.615:621.384.612.1† 1897
Pulsed Oscillator for F.M. Cyclotron—J. W. Burkig, E. L. Hubbard, and K. R. MacKenzie. (*Rev. Sci. Instr.*, vol. 20, p. 135; February, 1949.)

621.396.615:621.396.619.13 1898
Single-Valve Frequency-Modulated Oscillators: Parts 1 & 2—K. C. Johnson. (*Wireless World*, vol. 55, pp. 122–123 and 168–170; April and May, 1949.) Part 1: A coil in the anode circuit of a pentode, coupled inductively to the cathode circuit, is used to modulate the effective value of the main tuning inductance without disturbing the oscillatory circuit seriously. The amplitude of oscillation can be very nearly constant over frequency ranges at least as great as ± 15 per cent and at central frequencies up to 10 Mc, since all the effects of circuit capacitance can be tuned out and the tube input capacitance is unimportant.
Part 2: Details of design and operation.

621.396.615.17 1899
Theoretical and Experimental Study of a Generator of Periodic Slave-Frequency Electrical Pulses—R. Legros. (*Rev. Gén. Elec.*, vol. 58, pp. 143–154; April, 1949.) Two circuits are discussed, of which one is an amplifier-limiter transforming an input voltage of variable wave form, amplitude, and frequency into a square-wave voltage of the same frequency, but almost constant wave form and maximum amplitude. The second circuit is a pulse generator operated by this square-wave output and providing pulses of duration short compared with their recurrence period. Amplifier-limiter circuits using (a) diodes, (b) pentodes as limiters are investigated; neither diodes nor pentodes are entirely satisfactory by themselves, but a combination of the two gives excellent results. Pulses obtained by use of a *RC* differentiator circuit are shown to be practically independent, as regards peak voltage and pulse shape, of the initial voltage wave form applied to the amplifier limiter.

621.396.615.17:621.317.755 1900
A Linear Resistance-Charged Gas Relay Time Base—E. J. B. Willey. (*Electronic Eng.* (London), vol. 21, p. 101; March, 1949.) A circuit designed to avoid the type of transformer ordinarily used, which has one winding very well insulated and also screened to minimize the risk of pickup in the sensitive grid and cathode circuits of the relay. Full component details are given.

621.396.615.17:621.317.755 1901
A Wide-Range Saw-Tooth Generator—P. G. Sulzer. (*Rev. Sci. Instr.*, vol. 20, pp. 78–80; January, 1949.) Description of a circuit which can be used as a timebase at frequencies between 15 cps and 500 kc, with good linearity and rapid flyback. A cathode-coupled multivibrator is modified by including a parallel *RC* combination in the cathode lead of the first tube.

621.396.615.17:621.385.032.24 1902
Grid Current with RC Coupling—H. T. Ramsay. (*Wireless Eng.*, vol. 26, pp. 113–118; April, 1949.) A mathematical analysis of the effect of grid current on the performance of a cathode-coupled multivibrator (Schmitt) circuit, to find the optimum values for the components for operation at a given minimum frequency. Experimental results support the theory.

621.396.645 1903
High-Frequency Amplifier Stages, with Particularly High Sensitivity and Cut-Off Sharpness, for Short-Wave and Ultra-Short-Wave Apparatus—H. Rückert. (*Arch. Elek.* (Übertragung), vol. 3, pp. 24–31; January, 1949.) Discussion with particular reference to input-circuit design, noise factor, and the effect of feedback. Results with a circuit of high sensitivity are given and also a few data for some suitable pentodes.

621.396.645 1904
10-kw. F.M. Broadcast Amplifier—J. R. Boykin. (*FM-TV*, vol. 8, pp. 18–20; December, 1948.) A grounded-grid amplifier, whose anode voltage is only 3,700 volts. Two triodes, Type WL3X2500A-3, are used. An apparent efficiency of 100 per cent is achieved in the frequency band 88 to 108 Mc.

621.396.645:537.311.33:621.315.59 1905
Some Novel Circuits for the Three-Terminal Semiconductor Amplifier—W. M. Webster, E. Eberhard, and L. E. Barton. (*RCA Rev.*, vol. 10, pp. 5–16; March, 1949.) Equivalent circuits for dc and ac are proposed for a transistor. The ac circuit is a T-network of resistances between base, emitter (positive electrode), and collector (negative electrode) with a generator in series with the collector. The generator has an emf μ times the voltage developed across the emitter resistance by the emitter current. Three amplifier connections are analyzed. The first, already described by Bardeen and Brattain (264 of February) has the base as a common electrode, with input to the emitter and output from the collector. The two new connections, which have input and output impedances more suitable for most purposes, have the input to the base, and output respectively from the emitter and the collector. These connections are applied to a direct-coupled amplifier, a 2-terminal oscillator, a relaxation oscillator, and a flip-flop circuit.

621.396.645:539.16.08 1906
A Simple Counter-Tube Amplifier—H. Goldstein. (*Z. Angew. Phys.*, vol. 1, pp. 329–330; March, 1949.) A 2-tube amplifier suitable for operating a counting mechanism is described; its usefulness is limited by relatively low gain. A similar 3-tube circuit has adequate gain for most purposes. Both amplifiers are designed for 200-volt anode supply.

621.396.645:621.317.733 1907
An Amplifier for Use with Conductance Bridges—D. J. G. Ives and R. W. Pittman. (*Trans. Faraday Soc.*, vol. 44, pp. 644–646; September, 1948.) Two amplifying channels are used, one being fed with the bridge signal plus the inseparable interference and the other with interference alone, collected by an antenna near the bridge. The outputs are combined to cancel the interference, leaving a clear signal

in the final output. Full circuit details are given of an amplifier sensitive to a few microvolts at 1,200 to 5,000 cps.

621.396.645:621.383 1908
The Development of a Photoelectric A.C. Amplifier with A.C. Galvanometer—J. M. W. Milatz and N. Bloembergen. (*Physica, 's Grav.*, vol. 11, pp. 449–464; March, 1946. In English.) An ac galvanometer is fed by the same ac source as that for a synchronous motor driving a sector disk which interrupts the light falling on a photo cell. This device is sensitive only to 50-cps ac of the right phase. Fluctuations due to thermal effects were practically eliminated by cooling the photo cell with liquid air.

621.396.645.029.3 1909
Compact 6AS7G Amplifier for Residence, Audio Systems—C. G. McProud. (*Audio Eng.* vol. 33, pp. 17–19, 40 and 16–19; March and April, 1949.) Circuit and construction details of an amplifier having output power about 6 watts, switching for selecting radio or phonograph input, sufficient gain and low-frequency equalization for low-level magnetic pickups, separate high- and low-frequency tone controls, and remote control facilities.

621.396.645.371 1910
Negative-Feedback Amplifiers—C. F. Brockelsby. (*Wireless Eng.*, vol. 26, pp. 43–49; February, 1949.) When negatve feedback is applied to an amplifier with two or more stages the frequency characteristic develops peaks at the edges of the band if a certain small amount of feedback is exceeded. These peaks get higher and sharper with increasing feedback until self-oscillation begins. The flatness of the frequency characteristic can be improved, without using filters, by properly distributing the gain between the stages of the amplifier. The frequency characteristics of 2- and 3-stage amplifiers are discussed theoretically; results are confirmed experimentally. It is suggested that a 3-stage amplifier should have two identical wide-band stages and one narrow-band stage; this is contrary to the usual practice.

621.396.662 1911
Fixed-Frequency F.M. Tuners—F. A. Spindell. (*FM-TV*, vol. 9, pp. 16–17, 32; April, 1949.) A circuit diagram of the basic unit, model RP-23, is given and discussed. Models RP-24 and RP-25 have additional control units for selecting in sequence two preset audio volume levels and for switching on and off. The first half of a 12AX7 is associated with a bridged-T network and acts as a narrow-band amplifier at the control frequency. The second half of the 12AX7 amplifies the signal, which is applied to a 12AU7 operating a sensitive relay. This relay energizes a sequence relay which selects the required audio output level.

621.396.69:061.3 1912
Developments in Components—(*Wireless World*, vol. 55, pp. 133–138; April, 1949.) Brief details of exhibits at the Radio Component Manufacturers' Federation 1949 exhibition. See also *Overseas Eng.*, vol. 22, p. 310; April, 1949; and *Electronic Eng.*, vol. 21, pp. 146–149; April, 1949.

621.396.615.17/.18+621.317.35 1913
Waveform [Book Review]—Chance, Hughes, MacNichol, Sayre, and Williams. (*See* 2006.)

GENERAL PHYSICS

535+621.396 1914
Interference, Diffraction and Spectral Resolution in Optics and Radio—G. S. Gorelik. (*Uspekhi Fiz. Nauk*, pp. 407–415; November, 1948. In Russian.) A unified conception of the phenomena common to both sciences.

535.215:621.383 1915
On Photoelectric Voltages in Light-Absorbing Materials—A. P. Snoek and C. J. Gorter. (*Physica, 's Grav.*, vol. 11, pp. 426–432; February, 1946. In English.)

535.215:621.383 1916
On the Interpretation of Observations on the Photoelectric Voltages with Intermittent Light—C. J. Gorter, L. J. F. Broer, and A. P. Snoek. (*Physica, 's Grav.*, vol. 11, pp. 401–411; February, 1946. In English.)

537.521 1917
On the Dielectric Properties of a Gas Discharge—E. E. Salpeter and R. E. B. Makinson. (*Proc. Phys. Soc.*, vol. 62, pp. 180–188; March 1, 1949.) Formal expressions are derived for the space currents and electrode currents obtained by applying a sinusoidal voltage to a pair of electrodes projecting into a gas discharge.

537.521.7 1918
Experimental Investigations of the Electrical and Optical Phenomena in Spark Breakdown in Gases—E. Fünfer. (*Z. Angew. Phys.*, vol. 1, pp. 295–304; March, 1949.)

538.122 1919
Lines of Magnetic Force Made Visible—(*Radio Tech.* (Vienna), vol. 25, pp. 259–260; April, 1949.) A special gas-filled diode, when brought near the poles of a magnet, shows the lines of force as luminous streaks. The electrons inside the tube travel along a narrow path coincident with a tube of force and ionize the gas in the tube. The field can be plotted by moving the tube to different positions relative to the magnet poles. A tube with 8 anodes, giving 8 narrow electron beams, is described.

538.56:535.13:621.396.67 1920
The Radiation and Transmission Properties of a Pair of Semi-Infinite Parallel Plates: Part 1—A. E. Heins. (*Quart. Appl. Math.*, vol. 6, pp. 157–166; July, 1948.) A plane monochromatic electromagnetic wave is incident upon the plates, which are assumed thin and perfectly conducting. The amplitude and phase of the wave traveling between the plates are calculated. The structure is regarded as a two-dimensional receiving antenna and its radiation pattern is obtained. Its properties as a transmitting antenna will be considered later. See also 2756 and 3504 of 1947 (Carlson and Heins).

538.569.4:546.331.31–145.1 1921
Absorption of Ultra-High-Frequency Waves in Salt Solutions—S. K. Chatterjee and B. V. Sreekantan. (*Indian Jour. Phys.*, vol. 22, pp. 547–552; December, 1948.) For aqueous solutions of $MgCl_2$, $CuSO_4$ and KCl, the absorption maxima in the range 300 to 500 Mc shift toward higher concentration for higher frequencies. See also 3099 of 1948.

53/54 1922
Table of Physical and Chemical Constants [Book Review]—G. W. C. Kaye and T. H. Laby. Longmans, Green and Co., London, 10th edition 1948, 194 pp., 21 s. (*Proc. Phys. Soc.*, vol. 62, p. 209; March 1, 1949.) Considerable revision has been undertaken; the fundamental constants have been corrected in the light of recent determinations, and derived constants re-calculated.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.745:550.385 1923
Magnetic Storms and Solar Activity, 1948—H. W. Newton. (*Observatory*, vol. 69, pp. 38–40; February, 1949.) Provisional sunspot numbers, magnetic storms, and sudden commencements are tabulated and discussed.

523.746:550.38 1924
Observational Aspects of the Sunspot-Geomagnetic Storm Relationships—H. W. Newton. (*Mon. Not. R. Astr. Soc., Geophys. Supplement*, vol. 5, pp. 321–335; March, 1949.) Full paper, summary abstracted in 1370 June.

538.12:521.15 1
The Magnetic Field of Massive Rota Bodies—P. M. S. Blackett. (*Phil. Mag.*, 40, pp. 125–150; February, 1949.) Continua and elaboration of 3112 of 1947.

550.38 "1947.10/.12" 1
Cheltenham [Maryland] K—Indices October to December, 1947—In 379 of Ma the author's initials should be R. E.

550.384.4:551.510.535 1
Daily Magnetic Variations Near Equators—D. F. Martyn. (*Nature* (Lond vol. 163, pp. 685–686; April 30, 1949.) The served anomalies appear to require the ex ence in the ionosphere of a region of high e tric conductivity, about 15° of latitude w encircling the world approximately midv between the magnetic equator (line of zero clination) and the geomagnetic equator (e distant from the magnetic poles). See also 1 and 2790 of 1948 (Egedal) and 710 of A (Chapman).

551.510.52 1
The Structure of the Temperature Field Turbulent Stream—A. M. Obukhov. (*B Acad. Sci. U.R.S.S., Sér. Géogr. Géophys.*, 13, no. 1, pp. 58–69; 1949. In Russian.) Res obtained are mentioned in 2024 below.

551.510.535 1
On the Connections between an Oxyg Atom Zone, Ionized Layers and Radiation to the Mögel-Dellinger Effect—E. Schröer. *Met.*, vol. 1, pp. 110–113; January and F ruary, 1947.) Calculations based on the m probable values of dissociation and recombi tion coefficients indicate that the oxygen-at zone extends upward from a height of about km. Recombination takes place so slowly t the zone persists through the night. The *F*-la may be due to photoionization of the N_2 m cule. Radiation in the band 661 to 765Å co produce the *E*-layer by photoionization of radiation in the band 910 to 1010 Å cou similarly, produce a layer at a height of 60 100 km with ion concentration normally sufficient for the reflection of electromagne waves.

551.510.535 19
Ionospheric Disturbances and Their T restrial effects—G. Leithäuser. (*Funk Ton*, vol. 3, pp. 127–143; March, 1949.) general discussion, with particular reference observations in Germany of ionosphere la heights, aurora and sporadic-*E*, and correlati with propagation phenomena and geomagne effects.

551.510.535:523.3:621.396.11 19
Moon Echoes and Penetration of the Ion sphere—Kerr, Shain, and Higgins. (*See* 203

551.510.535:551.557 19
Atmospheric Currents at a Height of km—C. Hoffmeister. (*Z. Met.*, vol. 1, pp. 41; November and December, 1946.) Obser tions extending from 1922 to 1945 are tabulat and discussed. At heights above 100 km latitude 50°N, currents are observed in summ from SSW to SW with stable velocities of abo 50 miles per second and relatively little d persion in either direction or velocity. In win two current systems are found, one identi with that of summer but deviated sligh westward, the other consisting of polar-air c rents from NW to NE. Velocities are higher winter than in summer, with mean values 65 miles per second for the SW currents and miles per second for the northerly. The d persion in the observed values of direction a velocity is much greater in winter than in su mer. 21 per cent of the observed velocities ceeded 100 miles per second; velocities abo

miles per second were seldom noted; the est value was 267 miles per second. Possi- explanations of the effects are suggested.

510.535:621.317.79 1933
Sweep Frequency Ionosphere Equipment—er. (*See* 2004.)

510.535:621.396.11 1934
The Ionosphere over Mid-Germany in ruary 1949—Dieminger. (*Fernmelde- Z.*, vol. 2, p. 94; March, 1949.) Mean ies of the limiting frequencies and heights he *E* and *F* layers, deduced from measure- its at Lindau über Northeim, are plotted discussed. From February 12 to 21, the ting frequency of the F_2 layer was unex- edly high and on three occasions about mid- exceeded 14 Mc.

510.535:[621.396.61+621.396.621 1935
Ionosphere Sounding [equipment]—Maguer 2069.)

593.9 1936
The Luminescence of the Atmosphere—I. Khvostikov. (*Uspekhi Fiz. Nauk.*, no. 3, 372–386; November, 1948. In Russian.)

LOCATION AND AIDS TO NAVIGATION

396.9:523.3 1937
Detection of Radio Signals Reflected from Moon—J. H. DeWitt, Jr., and E. K. dola. (PROC I.R.E., vol. 37, pp. 229–242; rch, 1949.) The moon is considered as a ar target and formulas are obtained for the nuation between transmitter and receiver ennas. An experimental equipment and suring technique used for obtaining reflec- s from the moon are described and results ined are given. The possibility of using als reflected from the moon for communica- purposes is considered. See also 2915 of 8 (Grieg, Metzger, and Waer).

396.93 1938
V.H.F. Automatic Direction Finder—(*Engi-* (London), vol. 187, p. 197; February 18, 9.) A short account of the remotely operated allation at Brussels airport. Probable error uld be 1 degree or less, with a range of 100 tical miles to aircraft at 10,000 feet. The elopment of such equipment was fully dis- ed by Cleaver (1078 of May).

396.93:519.271 1939
Theory of Error Distribution: Application Radio Navigational Aids—P. F. Duncan. *reless Eng.*, vol. 26, pp. 49–52; February, 9.)

396.93.029.62 1940
Development of Single-Receiver Automatic ock Direction-Finders for Use in the Fre- ncy Band 100–150 Megacycles per Second F. Cleaver. (*Elec. Commun.*, vol. 25, pp. 362; December, 1948.) Reprint, with minor tions, of 1078 of May.

396.93.088 1941
H.F. Transmitter for D.F. Measurements G. Pressey. (*Wireless Eng.*, vol. 26, pp. 128; April, 1949.) Discussion of the princi- of design and the details of construction operation of an elevated screened-loop smitter. The loop can be rotated about a zontal axis to produce a wave of known rization for the measurement of the polari- n error of direction-finding systems. The llax error that occurs if a dipole is used as adiator is thus eliminated.

396.932:621.396.9 1942
The Development of Shipborne Naviga- l Radar—R. F. Hansford. (*Jour. Inst.* , vol. 1, pp. 118–141; April, 1948. Discus- pp. 141–147.) Survey based on two papers nted at the International Meeting whose rt was noted in 3135 of 1947. Some of the navigational data has been elaborated and the account has been brought up to date; possible future developments are discussed.

621.396.933 1943
Note on a Short-Range Radio Position-Finding System Using Modulated Continuous Waves—R. F. Cleaver. (*Elec. Commun.*, vol. 25, pp. 363–372; December, 1948.) Reprint, with minor additions, of 96 of February.

621.396.933.2 1944
Radio Distance-Measuring Equipment for Aerial Navigation—(*Engineering* (London), vol. 167, p. 187; February 25, 1949.) Long summary of paper abstracted in 1402 of June (Busignies).

621.396.933.2:621.396.826:621.396.619.16 1945
Elimination of Reflected Signal Effects in Pulsed Systems—D. O. Collup. (*Tele-Tech*, vol. 8, pp. 38–40, 64; January, 1949.) Multipath reflections can increase the apparent width of the shorter radar pulses in systems which depend on pulse width or pulse spacing to convey information. The incoming direct pulse can, however, be used to make the receiver insensitive to pulses whose amplitude is not constant over the whole pulse width. An echo-suppression circuit of this type for a radar transponder beacon is described.

621.396.933.24 1946
Rotating Field Beacons—F. de Fremery. (*Communication News*, vol. 9, pp. 2–20; September, 1947.) Beacons rotated mechanically cannot provide information quickly enough for antenna navigation. This difficulty can be overcome by rotating the field electrically, and comparing its phase with that of a pulsating field, but the two fields must not have the same frequency as they could not then be separated in the receiver. The properties of simple and modulated rotating fields are discussed. If two fields have harmonically related frequencies, the phase angle between them can be measured accurately with simple apparatus at the receiver, and this phase angle is a multiple of the azimuth. A rough bearing is obtained from modulation with fields of frequencies f_0 and $2f_0$ rotating in the same sense; the accurate bearing, which by itself would be ambiguous, is obtained from modulation of fields of frequencies f_0 and nf_0 rotating in opposite senses. Apparatus for transmitting and receiving such fields is described.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788 1947
Construction and Theoretical Analysis of a Direct-Reading Hot-Wire Vacuum Gauge with Zero Point Control—J. A. H. Kersten and H. Brinkman. (*Appl. Sci. Res.*, vol. A1, no. 4, pp. 289–305; 1949.) Description of a gauge connected to a special bridge, unaffected by voltage fluctuations or changes in ambient temperature. Pressures between 10^{-5} and 10^{-1} mm Hg can be measured.

531.788.7 1948
Design of an Ionization Manometer Tube—D. L. Hollway. (*Elec. Commun.*, vol. 25, pp. 373–385; December, 1948.) An abridged version was abstracted in 3916 of 1947.

535.37 1949
Luminescent Solids [Phosphors]—H. W. Leverenz. (*Science*, vol. 109, pp. 183–195; February 25, 1949.) A general discussion of the properties of typical phosphors, luminescence mechanism, and luminescence emission spectra.

535.37:535.61-15 1950
Inertia Effects in Infra-Red Sensitive Phosphors—F. R. Scott, R. H. Thompson, and R. T. Ellickson. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 64–67; January, 1949.) Discussion of results of measurements of the time required for certain phosphors to acquire maximum brightness when exposed to infrared radiation.

535.37:621.315.61.011.5 1951
The Optical and Electrical Properties of Zinc Silicate Phosphors—E. Nagy. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 42–49; January, 1949.) A close connection was found between dielectric loss and luminescence. See also 1051 of 1948 (Szigeti and Nagy).

537.311.33 1952
Electron Diffraction and Rectification from Silicon and Pyrite Surfaces—J. M. Cowley and J. L. Symonds. (*Trans. Faraday Soc.*, vol. 44, pp. 53–60; January and February, 1948.) For the best rectification, the crystal lattice should be almost perfect and free from fracture or mosaic structure.

538.213 1953
Permeability Decrease with Increasing Frequency—H. J. van Leeuwen. (*Physica, 's Grav.*, vol. 11, pp. 35–42; January, 1944. In Dutch, with French summary.) Theoretical discussion.

538.221 1954
The Magnetic Structure of High-Conductivity Alloys: Part 1—On Certain Peculiarities of the Magnetization Curves and Hystersis Loops of Alnico and Vicalloy—L. A. Shubina and Ya. S. Shur. (*Zh. Tech. Fiz.*, vol. 19, pp. 88–94; January, 1949. In Russian.) Part 2, 1955 below.

538.221 1955
The Magnetic Structure of High-Conductivity Alloys: Part 2—The Effect of Thermal Treatment on the Electrical Resistance of Alnico—V. I. Drozhzhina, M. G. Luzhinskaya, and Ya. S. Shur. (*Zh. Tekh. Fiz.*, vol. 19, pp. 95–99; January, 1949. In Russian.) Part 1, 1954 above.

538.221 1956
The Structure and Properties of the Alloy Cu_2MnIn—B. R. Coles, W. Hume-Rothery, and H. P. Myers. (*Proc. Roy. Soc.* A, vol. 196, pp. 125–133; February 22, 1949.)

538.221:621.317.4.029.62 1957
Ferromagnetism at Very High Frequencies: Part 2—Method of Measurement and Processes of Magnetization—Johnson and Rado. (*See* 1994.)

538.221:621.318.22:538.652 1958
The Cause of Anisotropy in Permanent Magnet Alloys—K. Hoselitz and M. McCaig. (*Proc. Phys. Soc.*, vol. 62, pp. 163–170; March 1, 1949.) Magnetostriction measurements on samples of Alcomax II indicate that, in the absence of a field, the domain magnetization is along the easy crystallographic direction which makes the smallest angle with the axis of anisotropy.

539.23:537.311 1959
Measurements of the Electrical Resistance of Thin Films of Copper, Silver and Lead—A. van Itterbeek and L. de Greve. (*Physica, 's Grav.*, vol. 11, pp. 78–90; February, 1944. In French.) Results are presented for films of thickness up to 100 mμ from room temperature down to the boiling point of liquid hydrogen. The temperature coefficient of resistance for Cu and Ag films is much less than that of the ordinary metal and becomes negative for a thickness of 3 mμ Cu or 4.5 mμ Ag.

539.23:537.311 1960
Measurements of the Electrical Resistance of Superposed Metallic Films—A. van Itterbeek and L. de Greve. (*Physica, 's Grav.*, vol. 11, pp. 465–469; March, 1946. In French, with English summary.) The resistance of Ag/Cu or Cu/Ag films agrees fairly well with that calculated for resistances in parallel, except when the film first deposited is extremely thin.

539.23:537.311:546.74 1961

Some Measurements on Thin Films of Nickel—A. van Itterbeek and L. de Greve. (*Physica, 's Grav.*, vol. 11, pp. 470–474; March 1946. In French, with English summary.) For the principal results see 343 and 3305 of 1946.

546.281.26 1962

On the Causes of the Nonlinearity of the Volt-Ampere Characteristic of Carborundum—V. I. Pruzhinina-Granovskaya. (*Zh. Tekh. Fiz.*, vol. 19, pp. 100–110; January, 1949. In Russian.)

546.431.82 1963

The Dielectric Behavior of $BaTiO_3$ Single-Domain Crystals—W. J. Merz. (*Phys. Rev.*, vol. 75, p. 687; February 15, 1949.)

549.514.51 1964

Production of Large Artificial Quartz Crystals—I. Franke and M. H. de Longchamp. (*Compt. Rend. Acad. Sci.* (Paris), vol. 228, pp. 1136–1137; March 28, 1949.) An autoclave process is described briefly. A seed crystal is arranged in a cool part and quartz or fused silica in the hottest part of the alkaline solution in the vessel, the pressure in which is maintained at 60 atmospheres and the temperature at 200 to 300°C. Homogeneous crystals $3\times2\times0.5$ cm, with good electrical properties, are produced in 3 to 4 weeks.

549.514.51:534.133 1965

Notes on the Frequency-Temperature Relationship of Some Low Frequency Quartz Plates—D. Fairweather and N. J. Beane. (*Marconi Rev.*, vol. 12, pp. 68–80; April to June, 1949.) Longitudinal oscillations are mainly considered; various types of cut are discussed. Flexural types of vibration are likely to increase in importance but more information is required. The Bell Telephone Laboratory terminology for angles of cut (see 1995 of 1944) is used. The constants controlling the form of the frequency versus temperature curves are derived by a method closely related to that of Mason (3518 of 1940). Design data are deduced from these and similar curves, and the relationships encountered in practice are discussed.

620.197 1966

The Climatization of Radio Equipment—M. A. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 115–120; April, 1949.) A note reviewing modern methods of protection against extremes of temperature, humidity, fungi and insects, together with a list of 70 recent references.

620.197:621.319.45 1967

On the Climatization of the Electrolytic Capacitor—H. E. Miquelis. (*Radio Tech. Dig.* (Franç.), vol. 3, pp. 77–82; April, 1949.) A few details are given concerning capacitors of several types; one type can withstand temperatures from $-60°$ to $+90°$C in store or during transit and functions satisfactorily in the range from $-40°$ to $+70°$C.

621.315.59:537.311.33:621.396.645 1968

Germanium—Important New Semiconductor—W. C. Dunlap, Jr. (*Gen. Elec. Rev.*, vol. 52, pp. 9–17; February, 1949.) An account of the chemical, physical, and electrical properties of Ge, and its application in the transistor.

621.315.59:621.319.4 1969

On the Use of Semiconducting Liquids for Impregnating Paper Capacitors—V. T. Renne. (*Zh. Tekh. Fiz.*, vol. 19, pp. 218–224; February, 1949. In Russian.)

621.315.61.011.5:546.431.82 1970

Anomalous Dielectric Properties of Polycrystalline Titanates of the Perovskite Type—J. R. Partington, G. V. Planer, and I. I. Boswell. (*Phil. Mag.*, vol. 40, pp. 157–175; February, 1949.)

621.315.612.011.5 1971

A New Type of Dielectric Polarization, and Losses in Polycrystalline Dielectrics—G. I. Skavani and A. I. Demeshina. (*Zh. Eksp. Teor. Fiz.*, vol. 19, pp. 3–17; January, 1949. In Russian.) Experiments were conducted with materials having a distorted crystal lattice. The materials were obtained by sintering TiO_2 with oxides of the second group of metals. Materials with small additions of SrO, CaO BaO, and ZnO have a high dielectric constant (of the order of 1,000) at frequencies from 10 to 20 kc. The loss angle is greater than for pure TiO_2, and has pronounced frequency and temperature maxima. Variations of ϵ and tan δ with increase of concentration of the alkaline earths are shown graphically. The experimental results are in agreement with a theory of relaxation polarization developed in earlier papers. The activation energy of loosely coupled ions and the frequency of their oscillations, as calculated from the experimental data, have abnormally low values.

621.315.618.015.5:537.52 1972

Breakdown Voltage of Rare-Gas/Nitrogen Mixtures between Hot Electrodes—J. A. M. van Liempt and W. D. van Wijk. (*Physica,'s Grav.*, vol. 11, pp. 167–178; March, 1944. In German.) The dependence of the breakdown voltage on the nitrogen content, the gas pressure, and the temperature and separation of the electrodes is shown graphically. The results are discussed with reference to the design of gas-filled glow lamps.

621.775.7 1973

Powder Metallurgy—G. Fitzgerald-Lee. (Electronic *Eng.*, vol. 21, pp. 87–90; March, 1949.) The development of the art is outlined and basic principles are briefly discussed. Methods of controlling grain growth are considered with particular reference to the production of tungsten wire with properties suitable for lamp filaments. Various applications are mentioned.

666.1.037.5 1974

The Technique of Glass-to-Metal Sealing with Special Reference to Vacuum-Tight Seals—A. G. Long. (*Jour. Soc. Glass Tech.*, vol. 30, no. 137, pp. 67–89; 1946.) A general survey ranging from wire seals used in tube pinches to cylindrical seals some 4 inches in diameter, from the standpoint of the author's personal experience. Practical manufacturing details are discussed and illustrated.

666.1.037.5:621.385.032.5 1975

The Electrode Leads of Transmitting Valves—E. G. Dorgelo. (*Communication News*, vol. 9, pp. 38–44; December, 1947.) Survey and discussion of various methods of sealing.

MATHEMATICS

517.53:621.392.52 1976

Splitting of an Analytical Function into an Even and an Odd Component, and also a Method of Determining, for a Known Even Component, the Odd Component and the Function as a Whole—Correction to 1422 of June, the title of which should read as above.

518.5 1977

Multiplication and Division by Electronic-Analogue Methods—E. M. Deeley and D. M. MacKay. (*Nature* (London), vol. 163, p. 650; April 23, 1949.) A multiplier should have (a) symmetry of response to positive and negative inputs, (b) an absolute indication of zero input, (c) independence of normal changes in electronic characteristics, and (d) rapid and accurate response. A multiplier having a cro with an axial magnetic field is here described; the transverse velocity of the electron stream is proportional to the voltage V_x applied to the X-plates of the cro. The magnetic field H produces a deflecting force initially in the direction, proportional to HV_x. This is coun acted by a photoelectric feedback system; spot can thus be held close to the X axis, the voltage V_y is then proportional to product of V_x and the current i_H producing field H. Conversely, the photo cell can be u to control H so that i_H is proportional V_y/V_x. Preliminary experimental results v this technique are satisfactory.

518.5:512.25 1

A Twelve-Equation Computing Inst ment—C. E. Berry and J. C. Pembert (*Instruments*, vol. 19, pp. 396–398; July, 19 An instrument of the decade-potentiom type for solving linear simultaneous equati by successive approximations.

518.5:517.944 1

A Note on Analog Computer Design—J. Bronzo and H. G. Cohen. (*Rev. Sci. Instr.*, 20, pp. 101–102; February, 1949.) In cert types of partial differential equations, spatial derivatives may be replaced by fi difference expressions. By using Laplace tr forms and matrices, the original equations be reformulated as a set of linear simultane equations. A simpler procedure sometimes sults from the application of a "simila transformation" to the matrix of these eq tions before a computer is designed to s them.

517.564.3(083.5) 1

Tables of Bessel Functions of Fractio Order. Vol. 1 [Book Review]—Natio Bureau of Standards. Columbia Univers Press, 1st edition 1948, 418 pp. (*Phil. M.* vol. 40, p. 124; January, 1949.) $J_\nu(x)$ is ta lated to 10 places of decimals for $\pm_\nu=\frac{1}{4},\frac{1}{2}$ and $\frac{3}{4}$, and $x=0(0.001)0.9(0.01)25.0$. Auxili tables are provided to extend the range values of x up to 30,000. See also 1708 of Ju

518.2 1

Practical Five-Figure Mathematical Tab [Book Review]—C. Attwood. Macmillan, L don, 1948. 74 pp., 3s. (*Wireless Eng.*, vol. p. 267; August, 1948; *Nature* (London), 163, p. 306; February 26, 1949.) The probl of the regions where mean proportional pa are untrustworthy has been resolutely tack by adjusting the interval of tabulation to s the rate of change of the function tabulat "The book . . . can be thoroughly reco mended."

MEASUREMENTS AND TEST GEAR

531.761+621.3.018.4(083.74) 19

Standard Frequency Broadcasts fr Hawaii—(*Tech. Bull. Nat. Bur. Stand.*, v 33, pp. 39–40; March, 1949.) An experimen station on the island of Maui, Territory Hawaii, now broadcasts standard frequenc (5, 10, and 15 Mc), time announcemer standard time intervals and standard musi pitch (440 cps) with call sign WWVH. Sim taneous reception of WWV and WWVH certain areas should not cause interference.

531.761+621.3.018.4(083.74) 19

The Atomic Clock—(*Tech. Bull. Nat. B Stand.*, vol. 33, pp. 17–24; February, 194 The operation depends upon the constant na ral frequency associated with the vibrations atoms in the NH_3 molecule. Accuracy is wit 1 part in 10^7; theoretical considerations in cate a potential accuracy within 1 part in or 10^{10}, according to the type of atomic syst and spectrum line used. The clock cons essentially of a 100 kc crystal oscillator frequency multiplier, a frequency discrimi tor, a frequency divider, a special 50-cps clo and a waveguide absorption cell contain NH_3 at a pressure of 10 to 15 μ Hg. The fun mental frequency of the oscillator is first mu

l up to 270 Mc by means of standard low-iency tubes. It is then multiplied up to) Mc by means of a frequency-multiplying tron, which is also modulated by a FM lator generating a signal at 13.8±0.12 The FM output is multiplied in a Si crystal fier to 23,870.4±0.96 Mc and fed to the nonia absorption cell. As the frequency of modulated control signal sweeps across absorption-line frequency (23,870.1 Mc) ne NH_3 vapor, the signal reaching the Si fier dips because of the absorption, giving gative output pulse. The output of the oscillator at 13.8±0.12 Mc is also fed to a iver together with a 12.5-Mc signal from quartz-crystal multiplying chain. When signal sweeps across the proper frequency 5 Mc+the 1.39 Mc if of the receiver) a nd output pulse is generated. If the time val between these two pulses is incorrect, ntrol signal is generated in a discriminator iit, and fed to a reactance tube which forces quartz-crystal circuit to oscillate at the ect frequency. See also *Radio-Electronics*, 20, pp. 74–76; March, 1949; and 1712 of (Huston and Lyons).

317.2:621.397.6 1984
TV Distribution System for Laboratory—J. Fisher. (*Communications*, vol. 29, pp. 43; February, 1949.) Description of a ralized system for producing the standard A composite video signal and distributing a number of laboratories with minimum ortion. Sources of signal include (a) local vision broadcasting stations, (b) monoscope al, and (c) picture signal from a crt flying-scanner.

317.3+621.317.7]:061.3 1985
Papers Digested for Conference on High-uency Measurements—(*Elec. Eng.*, vol. pp. 251–257; March, 1949.) Authors' sum-es of most of the papers read at the con-ice.

317.3:621.395.44 1986
Maintenance Measurements on Carrier phony Equipment—J. de Jong. (*Philips . Rev.*, vol. 8, pp. 249–256; August, 1946.) ussion of apparatus used and operational irements.

317.324†:621.318.4 1987
Coils as H.F. Measurement Probes for Ab-e Field-Strength Determinations—E. Ro-en. (*Funk und Ton*, vol. 3, pp. 167–172; ch, 1949.) Measurements with 5 different s of coil indicate that a small single-layer drical coil is particularly suitable for such lute measurements, since its effective area be determined accurately from its mean netrical diameter.

317.329:538.122 1988
Electrolyte] Tank Model for Magnetic lems of Axial Symmetry—R. E. Peierls T. H. R. Skyrme. (*Phil. Mag.*, vol. 40, 69–273; March, 1949.) Magnetic problems involve vortices and cannot, therefore, rectly represented in the electrolyte tank. 2-dimensional problems, the conjugate can be studied; this technique is here ed to systems which are very nearly plane have axial symmetry.

317.335.3†+621.317.374 1989
n Optical Method for Measuring the Di-ric Constant and Dielectric Losses of Solid ctrics in the Centimetre Wavelength ge—L. L. Odynets. (*Zh. Tekh. Fiz.*, vol. p. 120–125; January, 1949. In Russian.) retical discussion of a method based on urements of the transparency of a slab of ielectric, with a description of the appara-Fig. 1) used for measuring the trans-ncy of glass and ebonite for $\lambda=5.5$ cm. method would be even more suitable for shorter wavelengths since the required over-all dimensions of the measuring apparatus would be smaller; liquid dielectrics could also be used.

621.317.34:621.315.1/.2 1990
No-Load and Short-Circuit Measurements for Determining the Transmission Characteristics of Open Lines and Cables—O. Naumann. (*Arch. Tech.* (Messen), pp. T22–T23; March, 1949.) The approximate methods used for calculating the low-frequency transmission characteristics of cables from no-load and short-circuit measurements of voltage, current, and power are not applicable at high frequency. More accurate methods given by Kaden (1063 of 1937), by Sommer for very low frequencies (4070 of 1939), and by Goldschmidt for high-frequency (1724 of 1943) are discussed.

621.317.353:621.396.619.13 1991
The Determination of the Distortion in a Frequency-Modulator—F. L. H. M. Stumpers and W. W. Boelens. (*Communication News*, vol. 9, pp. 107–109; August, 1948.) As the modulation voltage is increased, the amplitudes of the components of the output frequency spectrum vary and pass through the value zero. The zero points can be used for measuring frequency deviation. The displacement of the zeros, caused by nonlinearity of the modulation characteristic, can be used for the measurement of odd harmonics. A shift in frequency of the output spectrum as a whole can be used to measure even harmonics.

621.317.372 1992
Improved Accuracy with a "Q"-Meter by the Use of Auxiliary Components—A. C. Lynch. (*Electronic Eng.* (London), vol. 21, pp. 91–93; March, 1949.) Accuracy can be considerably increased by using a mirror galvanometer in parallel with the meter giving Q values, a second galvanometer, suitably shunted, in series with the meter measuring the resonant-circuit response and a variable capacitor with a fine scale of high calibration accuracy in parallel with the fitted tuning capacitor. The principle used is that of the reactance-variation method described by Hartshorn and Ward (351 of 1937), but here the input voltage is varied in known ratios, and the voltage in the circuit is brought to a fixed value by detuning. A circuit diagram is given. A Q-meter using this principle is described; accuracy is discussed, and possible methods of further improvement are suggested.

621.317.39.029.6:536.33 1993
Conditions for Maximum Sensitivity of U.H.F. Radiometers—J. L. Steinberg. (*Onde Élec.*, vol. 29, pp. 160–166; April, 1949.) Discussion of apparatus of the type described by Dicke (475 of 1947). With optimum noise factor and input circuit, and a certain amount of negative feedback, it is possible to measure the apparent temperature of a receiving antenna to within about 0.3°C. Apparatus designed for investigation of 1,200-Mc radiation, constructed at the physical laboratory of the École Normale Supérieure, uses a dipole-fed parabolic reflector 3 meters in diameter mounted on an equatorial support. The superheterodyne receiver has a bandwidth of 20 Mc and a noise factor of 3.5. Further details of this equipment will be published later.

621.317.4.029.62:538.221 1994
Ferromagnetism at Very High Frequencies; Part 2—Method of Measurement and Processes of Magnetization—M. H. Johnson and G. T. Rado. (*Phys. Rev.*, vol. 75, pp. 841–864; March 1, 1949.) Part 1: 3182 of 1947 (Johnson, Rado, and Maloof.)

621.317.7.029.63/.64(083.74) 1995
Microwave Measurement Standards—(*Jour. Frank. Inst.*, vol. 247, pp. 156–161; February, 1949.) Discussion of standards and calibration services available or methods being developed at the National Bureau of Standards for frequencies from 300 Mc to over 100,000 Mc. These include: (a) a frequency standard for 300 to 40,000 Mc accurate within 1 part in 10^8, (b) methods and equipment for dielectric measurements at frequencies near 1,000, 3,000, 9,000, and 24,000 Mc, (c) a primary standard of attenuation consisting of a waveguide operating at a frequency below cutoff, which can be used for any microwave frequency by means of a development of the if substitution method, (d) a microwave power-measuring console nearly completed, (e) investigation of the accuracy of bolometers and thermal noise sources as power standards, (f) development of primary and secondary frequency standards using spectrum lines of gases, (g) a prototype atomic clock (see also 1983 above), and (h) measurement of spectrum lines as secondary frequency standards.

621.317.71/.72 1996
An Instrument for the Measurement and Time Integration of Small Voltages and Currents—I. A. D. Lewis and A. C. Clark. (*Jour. Sci. Instr.*, vol. 26, pp. 80–84; March, 1949.) Description with full component details. Potentials varying between 0 and 0.5 volt and their time integrals can be measured within 3 per cent. Currents can be measured with similar accuracy; full-scale deflection on the lowest range is 0.01 μamp.

621.317.71:621.385 1997
Reduction of Noise in Thermionic Electrometers with Mechanical Conversion—H. den Hartog and F. A. Muller. (*Physica,'s Grav.*, vol. 11, pp. 161–166; March, 1944. In Dutch, with English summary.) Continuation of 858 of 1945.

621.317.715.004.64:538.22 1998
Non-Ferrous Copper Wire for Moving-Coil Meters—P. G. Moerel and A. Rademakers. (*Philips Tech. Rev.*, vol. 8, pp. 315–319; October, 1946.) Effects due to Fe impurity in the Cu wire of the coils is discussed and methods of producing wire with extremely low Fe content are described.

621.317.738 1999
A Note on the Measurement of Four-Terminal Inductances by Astbury's Method—L. H. Ford. (*Jour. Sci. Instr.*, vol. 26, pp. 108–109; March, 1949.) Astbury's method (noted in 1932 Abstracts, p. 48) is unsuitable in its original form for frequencies above af, but if the fixed resistance ratio arms of the bridge are replaced by a Kelvin-Varley potential divider of known phase defect, the method can be used for frequencies up to 100 kc. Experimental results on a 10-μh coil show good agreement at all frequencies from 1 to 100 kc with values obtained by other methods.

621.317.755 2000
A 3-Beam Micro-Oscillograph for Display of Oscillations up to 10,000 Mc/s—v. Fe. (*Frequenz*, vol. 3, pp. 19–22; January, 1949.) A short description of Lee's instrument (1692 of 1946) with a tabular comparison with the Rogowski, Siemens, and A.E.G. oscillographs.

621.317.761 2001
Heterodyne Frequency Meter for High Frequencies—L. Liot. (*Radio Franç.*, pp. 14–17; April, 1949.) Description, with circuit details, of an instrument for the rapid measurement of frequencies from 5 to 1,000 Mc. The local oscillator, which is of the Lecher-line type and uses a Type 955 triode, has a frequency range of 69 to 220 Mc; within this range unknown frequencies are measured by a direct zero-beat method, while harmonic methods are used outside the range.

621.317.772.029.54/.58:621.396.645.37 2002
A Phase Meter for the Frequency Band 100 kc/s–20 Mc/s—W. T. Duerdoth. (*P.O. Elec. Eng. Jour.*, vol. 42, part 1, pp. 43–46; April, 1949.) Intended primarily for measuring the phase change round the feedback loops of amplifiers. An accuracy within $\pm 3°$ is possible provided that the loop gain or loss does not exceed 40 db. The magnitudes of the voltages whose phases are to be compared are made equal by means of two special variable-gain amplifiers which cause the same phase change in received signals whatever their respective gains. The design of these amplifiers and of the necessary frequency changer is discussed and circuit diagrams are given.

621.317.784 2003
Pulse Power Measurement by a Heterodyne Method—L. S. Schwartz. (*Communications*, vol. 29, pp. 26–27; February, 1949.) Pulse width and repetition rate need not be known. The rf pulse and cw oscillations of nearly the same frequency are applied to the square-law detector of a synchroscope receiver. The cw is adjusted so that the peak value of the variable component of the detection voltage equals the amplitude of the envelope of the rf pulse. The rf pulse power is then 6 db above the cw power at the point of entrance into the detector. Sources of error are discussed; accuracy is within a few per cent.

621.317.79:551.510.535 2004
Sweep Frequency Ionosphere Equipment—P. G. Sulzer. (*Jour. Appl. Phys.*, vol. 20, pp. 187–196; February, 1949.) The equipment records ionosphere virtual height as a function of frequency over the range 1 to 25 Mc. Special features of the device are high power output, good receiver sensitivity, and anti-jamming circuits. Detailed circuit diagrams are included. See also 2240 of 1948 (Thomas and Chalmers).

621.317.791 2005
A Self-Checking Wobbulator—J. H. Vogelman. (*Communications*, vol. 29, pp. 28–31; February, 1949.) A portable test set for frequencies between 5 and 100 Mc, comprising a FM signal generator, frequency meter, and cro which can be used to measure the gain, frequency, bandwidth, and tuning characteristics of wide-band if and rf amplifiers and receivers.

621.317.35+621.396.615.17/.18 2006
Waveforms [Book Review]—B. Chance, V. Hughes, E. F. MacNichol, D. Sayre, and F. C. Williams (Eds). McGraw-Hill, London, 785 pp., £3. (*Wireless Eng.*, vol. 26, p. 139; April, 1949.) Vol. 19 of the MIT Radiation Laboratory series. The book is more descriptive than analytic and deals mainly with nonlinear circuits. The action of electronic switches, frequency multipliers and dividers, and counting circuits is discussed.

621.317.35+621.396.619+621.396.822 2007
Frequency Analysis, Modulation and Noise [Book Review]—S. Goldman. McGraw-Hill, London, 1948, 434 pp., 36s. (*Electronic Eng.* (London), vol. 21, p. 152; April, 1949; PROC. I.R.E., vol. 37, p. 541; May, 1949.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.61–15:621.383 2008
Direct Recording of Spectra in the Region 1.2 μ to 3 μ using the Lead Sulfide Photo-Conductive Cell—R. C. Nelson. (*Jour. Opt. Soc. Amer.*, vol. 39, pp. 68–71; January, 1949.) See also 3330 of 1947 (Cashman).

621.317.39 2009
Electronic Gauges—J. Schwartz. (*Microtecnic* (Lausanne), vol. 3, pp. 10–18; January and February, 1949. In English.) Conclusion of 1736 of July.

621.365.5+621.365.92 2010
High-Frequency Heating—S. W. Scherer. (*Communication News*, vol. 9, pp. 45–55; December, 1947.) Discussion of both induction and dielectric heating, and of their applications. Several examples are given of apparatus designed for particular applications.

621.38.001.8 2011
The Electronic Brain—W. R. Ashby. (*Radio-Electronics*, vol. 20, pp. 77–80; March, 1949.) Reprint of 1144 of May.

621.38.001.8 2012
Electronics in the Service of Industry—(*Radio Tech. Dig.* (Franç.), vol. 3, pp. 105–111; April, 1949. Bibliography, pp. 111–113.) Brief general discussion, with some details of photoelectric counters and opacity meters.

621.38.001.8:061.3 2013
Papers Digested for Conference on Electronic Instrumentation—(*Elec. Eng.*, vol. 68, pp. 246–251; March, 1949.) Authors' summaries of most of the papers read at the conference.

621.384.611.1† 2014
A 20-MeV Betatron—W. Bosley, J. D. Craggs, D. H. McEwan, and J. F. Smee. (*Proc. IEE*, part I, vol. 96, pp. 85–86; March, 1949.) Discussion on 179 of February.

621.384.611.1† 2015
A New Type of [9-MeV] Betatron without an Iron Yoke—A. Bierman. (*Nature* (London), vol. 163, pp. 649–650; April 23, 1949.) Two coils are used in series, with a sealed-off glass acceleration tube between them. The dimensions are chosen so that the magnetic field at a point distant r from the center is proportional to r^{-n}, where $0<n<1$. The flux required within the electron orbit is obtained by means of a small iron core along the axis. The current through the coils is obtained by periodical discharges of a 6.5 -$\mu\mu$f capacitor across a spark gap; these discharges occur every few seconds and initiate damped 2.5-kc oscillations, with a peak current of about 5,000 amperes. The whole betatron only weighs about 50 kg and the simple construction makes the cost low.

621.384.612.1† 2016
Design of the Radiofrequency System for the 184-Inch Cyclotron—K. R. MacKenzie, F. H. Schmidt, J. R. Woodyard, and L. F. Wouters. (*Rev. Sci. Instr.*, vol. 20, pp. 126–133; February, 1949.) See also 1712 of 1948.

621.384.612.1†:621.396.615 2017
Pulsed Oscillator for F.M. Cyclotron—J. W. Burkig, E. L. Hubbard and K. R. MacKenzie. (*Rev. Sci. Instr.*, vol. 20, p. 135; February, 1949.)

621.384.621.1† 2018
The High-Voltage Electrostatic Generator at the Atomic Energy Research Establishment—R. L. Fortescue and P. D. Hall. (*Proc. IEE*, part I, vol. 96, pp. 77–85; March, 1949.) A 5-Mv generator of the pressurized Van de Graaf type.

621.385.833 2019
Aberration Correction with Electron Mirrors—E. G. Ramberg. (*Jour. Appl. Phys.*, vol. 20, pp. 183–186; February, 1949.) Formulas for spherical and chromatic aberration are applied to a concave electron mirror with concentrated field distribution. The aberration coefficients of such mirrors are so large that this method of correction has serious practical difficulties.

621.396.615.17:615.849 2020
Electromedical Stimulators—O. B. Sneath and E. G. Mayer. (*Wireless World*, vol. 55, pp. 129–132; April, 1949.) Pulses are required of lengths between 1 second and 10 μseconds, with repetition rates between 1 and 50 per second and maximum output voltage of the order of 100 volts at 100 mamp. Circuit producing such pulses are discussed.

621.398
Radio Control of Mobile Miniatures—L. Safford, Jr. (*CQ.*, vol. 5, pp. 18–21; April, 1949.) A system for starting, stopping and steering model boats or cars.

PROPAGATION OF WAVES

538.566
Unification of the Formulae Represe the Principle of Huyghens for Electromag Waves—F. Croze and G. Darmois. (*C Rend. Acad. Sci.* (Paris), vol. 228, pp. 826; March 7, 1949.) Three conditions mu satisfied if a system of formulas is to physically and mathematically correct pression of Huyghens' principle. The form proposed successively by Love, Macdo Larmor, Bromwich, Schelkunoff, L. de Bro Novobatzky (for a finite surface), and F all satisfy these three conditions; in co quence, they can all be referred to a com form, which is here given. The condition the secondary waves from the various elem of the surface considered must be pure ele magnetic waves is not satisfied by the form of Kirchhoff, nor by those proposed success by Ignatowsky, Tonolo, Tedone, Ko Stratton, and Chu and (for a closed sur Novobatzky. This is shown by considering case where the surface in question is a s surface, either plane or of large radius. also 1462 and 1463 of June.

538.566.2
On the Propagation of Waves in an homogeneous Medium—O. E. H. Rydb (*Chalmers Tekn. Högsk. Handl.*, no. 74, 3 1948. In English.) A mathematical theor propagation in an inhomogeneous or strat medium is developed to determine the limi conditions in which more approximate the are applicable. For a slightly inhomogen medium, first- and higher-order approximat to the wave equations are obtained and t usefulness is estimated by application to sp cases for which exact independent solut exist. Large variations in refractive index then considered. The results are comp with those of other methods. The theor applied to the propagation of magneto-hy dynamic waves in the sun and to the propagation of radio waves in the lo troposphere.

621.396.11
The Effect of Pulsations of the Refrac Index of the Atmosphere on the Propaga of Ultra-Short Waves—V. A. Krasil'ni (*Bull. Acad. Sci. U.R.S.S., sér. Géogr. Géoph* vol. 13, no. 1, pp. 33–57; 1949. In Russi The usual interpretation of fading as a re of interference between the direct and indi rays cannot be strictly correct, especially w transmission extends beyond the horizon. suggested that pulsations of the refractive dex of the atmosphere due to tempera fluctuations may have an important bea on fading. It is shown mathematically when the scale of pulsations is smaller than wavelength, the waves are dispersed and w it is greater than the wavelength, path va tions occur which lead to variations in int sity at the receiving point. Using the meth of geometrical optics and taking into acco the results obtained by A. M. Obukhov (1 above) formulas are derived for determin fluctuations of the amplitude and phase at point of reception. Experimental results av able in the literature are discussed in de and are in general agreement with the prope theory. It is concluded that: (a) when tr mission is within the horizon, usw fading be explained by the temperature fluctuati of the atmosphere; (b) fading when transm sion is somewhat beyond the horizon is ma caused by the interference between the

onents but it is also necessary to take into nt the effect of the temperature fluctua- on both components; (c) at very great nces, when reception is carried out only e indirect ray, the temperature-fluctuation again becomes predominant.

96.11:551.5 2025
n Extension of Macfarlane's Method of cing Refractive Index from Radio Obser- ns—A. W. Straiton. (*Jour. Appl. Phys.*, 20, p. 228; February, 1949.) Comment on of 1947. The second set of height-gain urements there suggested can be replaced set of height versus phase measurements; able method of measuring phase was dis- d in 3225 of 1948 (Straiton and Ger-).

96.11:551.510.535 2026
bsorption of Radio Waves Reflected at cal Incidence as a Function of the Sun's h Angle—E. W. Taylor. (*Jour. Res. Bur. Stand.*, vol. 41, pp. 575–579; Decem- 948.) Analysis of Central Radio Propaga- Laboratory records, extending over 3 , indicates that absorption depends ap- mately linearly on the cosine of the sun's n angle.

96.11:551.510.535 2027
Note on the Ionospheric Absorption lem—L. G. McCracken. (*Jour. Appl.* ., vol. 20, pp. 229–230; February, 1949.) ormula for the total nondeviating *E*-layer ption derived by Best and Ratcliffe of 1938) is here obtained without using in of their approximations. Appleton of 1938) obtained a similar formula.

96.11:551.510.535 2028
he Ionosphere over Mid-Germany in uary 1949—Dieminger. (*See* 1934.)

96.11:551.510.535 2029
hanges in Radio Reception during Sun- Period '45–47—H. T. Stetson. (*Tele-Tech*, , pp. 29, 73; December, 1948.) Summary mer. Astr. Soc. paper. Variations of F_2 al frequencies and of field intensities at 1 10 Mc with variations of sunspot ac- are recorded for the recent rise period. nal field-intensity changes to be expected g the decline period are discussed.

96.11:551.510.535:523.3 2030
Moon Echoes and Penetration of the sphere—F. J. Kerr, C. A. Shain, and C. S. ns. (*Nature* (London), vol. 163, pp. 310– February 26, 1949.) The possibility of reflections from the moon to extend the of the ionosphere is investigated. Stations and VLB5 operating at frequencies near c were used. The equipment and the ex- ental procedure are described. Echoes obtained in 13 out of 15 trials. Wide tions in the amplitude of echoes were re- d; on no occasion did the amplitude ex- the theoretical value. Fading periods of nd or less were observed, similar to those ienced with ionospheric reflections. Echo encies were about 50 cps higher than those e transmitted signal, mainly because of Doppler effect of the earth's rotation. As noon rose, echoes could not be obtained n as could be expected from calculations on vertical-incidence ionospheric sound-

nalysis of 20-Mc solar noise also shows pheric effects, with the significant dif- e that solar noise energy is received at ectedly low elevations of the sun.

96.812.029.62 2031
n Normal and Abnormal Refraction of -Short Waves in the Atmosphere—H. H. er. (*Z. Met.*, vol. 2, p. 86; March, 1948.) -strength records of usw transmissions .3 meters) over a 63-km path, about 40 nt beyond the optical range, are discussed briefly. The considerable increases of field strength occasionally observed are accompanied by relatively little fluctuation and can be attributed to certain local characteristics of the lower atmosphere.

621.396.812.029.64 2032
Effect of the Atmosphere on Microwaves—H. H. Klinger. (*Z. Met.*, vol. 2, pp. 314–316; October, 1948.) Measurements of received signal strength for $\lambda = 1.36$ cm showed strong absorption due to rain or even a slightly damp atmosphere, whereas 3.6-cm waves under the same conditions were not appreciably affected. Further investigations with wavelengths of 1 to 10 mm are proposed.

621.396.812.029.64 2033
Low-Level Atmospheric Ducts—R. F. Jones, J. S. McPetrie and B. Starnecki. (*Nature* (London), vol. 163, p. 639; April 23, 1949.) Comment on 1167 of May. Jones explains the presence of the ducts observed both in cold and warm weather in terms of the previous history of the air at heights between 200 and 2,000 feet. McPetrie and Starnecki consider that although Jones' hypothesis is satisfactory for periods of high wind-velocity, low air temperature rather than high wind-velocity is the main cause of the cold-weather ducts.

621.396.812.029.64 2034
Oversea Propagation on Wavelengths of 3 and 9 centimeters—J. S. McPetrie, B. Starnecki, H. Jarkowski, and L. Sicinski. (Proc. I.R.E., vol. 37, pp. 243–257; March, 1949.) For other accounts of results obtained at the same sites see 518 of 1947 (Megaw) and 2329 of 1948 (McPetrie and Starnecki).

621.396.812.3 2035
A Peculiar Type of Rapid Fading in Radio Reception—N. S. Subba Rao and Y. V. Somayajulu. (*Nature* (London), vol. 163, p. 442; March 19, 1949.) Discussion of the "flutter phenomenon," a variation in intensity observed at Waltair, India, when receiving broadcast transmissions on wavelengths of 41 and 60 meters. The effect is observed occasionally on a wavelength of 19 meters but is completely absent at medium frequency. The flutter occurs only during the hot season (February to June) and begins about sunset; the fluctuation frequency is very low (0.3 to 0.5 cps) before sunset, and increases to 2 to 2.4 cps early in the night, thereafter remaining constant for long periods. The ionosphere must be the cause of this phenomenon, which may be due to the rapid movement of ionic clouds across the *F* layer. See 3279 of 1946 (Wells, Watts, and George).

621.396.812.3:551.510.535 2036
Short-Range Fading of Broadcasting Transmissions—W. Gerber and A. Werthmüller. (*Tech. Mitt. Schweiz. Telegr. Teleph. Verw.*, vol. 25, pp. 1–12; February 1, 1947. In German.) The reflecting properties of the ionosphere in the medium-wave band are discussed. The fine structure of the fading diagram is attributed to interference effects, while the main features are principally determined by absorption and are related both to the sunspot period and to the season, with maximum fading effects in spring and autumn and minimum occurrence in summer and winter. The solar effects are superposed on the seasonal effects and are greatest at sunspot minimum and least at sunspot maximum. Curves are given showing the average fading effects for the Beromünster transmitter as observed at St. Gallen, at different times during the evening, from April, 1936, to November, 1946. The curves all show maxima about 1944, with subsidiary peaks about 1941 and minima near 1938.

621.396.812.3.029.54 2037
Space-Wave Absorption and Large-Scale Weather Conditions—G. Falckenberg and E. Lauter. (*Z. Met.*, vol. 2, pp. 259–265; September, 1948. Transmissions from Kalundborg on $\lambda = 1,250$ meters were observed at Warnemünde 180 km distant. In the daytime, the transmissions were almost completely absorbed by the ions in the lower part of the ionosphere, but at night the absorption showed wide variations related to projections into the ionosphere of equatorial or polar air masses. A connection between the absorption and air-pressure variations at a height of 9 km was also established.

RECEPTION

621.396.619.13:621.392 2038
Distortion of Frequency-Modulated Signals in Electrical Networks—F. L. H. M. Stumpers. (*Communication News*, vol. 9, pp. 82–92; April, 1948.) Long summary of part of the thesis noted in 2221 of 1947. See also 1886 of 1948.

621.396.621+621.396.61]:551.510.535 2039
Ionosphere Sounding [equipment]—Maguer. (*See* 2069.)

621.396.621:621.392.52:621.396.5 2040
Application of Crystal Filters in [telegraphy] Receivers—Maarleveld. (*See* 1892.)

621.396.621:621.396.5 2041
A Modern Receiver for Radiotelegraphy—C. T. F. van der Wyck. (*Tijdschr. ned. Radiogenoot.*, vol. 14, pp. 27–39; March, 1949. Discussion, p. 40. In Dutch, with English summary.) Description, with block diagram, of the receiver, and discussion of (a) considerations leading to its design, (b) automatic tuning control, and (c) conditions for a stable circuit. See also 3510 of 1948 and 1892 above.

621.396.81:621.396.9 2042
Signal/Noise Ratio in Radar—S. de Walden. (*Wireless Eng.*, vol. 26, pp. 140–141; April, 1949. Comment on 2899 of 1948 (Levy).

621.396.826:621.396.933.2:621.396.619.16 2043
Elimination of Reflected Signal Effects in Pulsed Systems—Collup. (*See* 1945.)

621.396.822+621.317.35+621.396.619 2044
Frequency Analysis, Modulation and Noise [Book Review]—Goldman. (*See* 2007.)

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2045
Evaluation of Transmission Efficiency According to Hartley's Expression of Information Content—A. G. Clavier. (*Elec. Commun.*, vol. 25, pp. 414–420; December, 1948.) Transmission efficiency, defined by extending Hartley's expression for telegraphic signals to telephony in presence of noise, is calculated for the main pulse-transmission systems. The resulting expressions enable the various systems to be compared, although simplicity and cost of equipment will also be important factors for deciding their relative merits. See also 515 of March (Landon) and 1361 of June (Shannon).

621.395.43/.44 2046
Study of the General Characteristics of the L. T. T. 12-Channel Carrier-Current Systems for Overhead Lines—H. Pech. (*Câbles and Trans.* (Paris), vol. 3, pp. 177–193; April, 1949.) General problems in connection with such systems are discussed, including (a) attenuation within the prescribed transmission band, (b) characteristic impedance of actual lines, (c) crosstalk, and (d) noise. A description is given of equipment constructed for the French Post Office. The frequency band covered is 30 to 150 kc. Details are included of the modulation and demodulation processes, terminal equipment, filters, equalizers and carrier-frequency generators. Test results on prototype equipment will be published later.

621.396.1:621.397.5 2047

Allocation of Frequencies for the Television Service—F. C. McL. (*BBC Quart.*, vol. 4, pp. 54–56; April, 1949.) The frequency band at present allocated is 41 to 66.5 Mc; it is hoped that the upper limit will shortly be raised to 68 Mc. For all new stations, asymmetric-sideband transmission will be used to enable 5 exclusive channels to be obtained in the full band, but transmissions from Alexandra Palace will remain unchanged.

621.396.619.13/.14:534.78 2048

Ratio of Frequency Swing to Phase Swing in Phase- and Frequency-Modulation Systems Transmitting Speech—D. K. Gannett and W. R. Young. (PROC. I.R.E., vol. 37, pp. 258–263; March, 1949.)Theoretical and experimental results are discussed. The ratio was found to vary with different voices, with the microphone and circuit characteristics, and with the kind of volume regulation used.

621.396.65.029.63 2049

Choice of Suitable Heights [of stations], Distance and Wavelengths in Planning Decimetre-Wave Links—A. Grün. (*Fernmeldetech. Z.*, vol. 2, pp. 69–72; March, 1949.) In the case of decimeter-wave systems, fading is due chiefly to interference between the direct ray and that reflected from an intermediate point of the earth's surface. Calculations show that in order to obtain as great a received field strength as possible, with little fading, the wavelength for a given distance between stations should not be too small and optimum heights should be chosen for the antennas. Formulas are given from which the optimum heights can be calculated, with curves for ranges of 50 and 60 km respectively and wavelengths from 10 to 100 cm.

621.396.712.2 2050

WMGM Master Control Equipment Design—M. E. Gunn. (*Audio Eng.*, vol. 33, pp. 24–28, 40; March, 1949.) Details, including block diagram, of a high-power broadcast station installation.

621.396.9:523.3 2051

Detection of Radio Signals Reflected from the Moon—DeWitt and Stodola. (*See* 1937.)

621.396.619+621.317.35+621.396.822 2052

Frequency Analysis, Modulation and Noise [Book Review]—Goldman. (*See* 2007.)

SUBSIDIARY APPARATUS

621.316.722.077.65:621.396.68 2053

Commutation in Rectifiers Using Relay Tubes: Parts 1 and 2—T. Douma. (*Communication News*, vol. 9, pp. 70–81 and 110–120; April and August, 1948.) A detailed discussion of the behavior of the well-known 3-phase Graetz circuit, taking account of the effect of leakage inductance and self-capacitance in the power transformer. Means of protecting this transformer from steep-fronted voltage surges are also indicated. The control characteristic of grid-controlled rectifiers is discussed in an appendix.

621.316.726 2054

Frequency Correction Equipment for Railway Signalling Supplies—(*Engineer* (London), vol. 187, pp. 184–186; February 18, 1949.) A 10-kva induction motor has a 440-volt three-phase stator and a single-phase rotor, the speed of which is controlled by a frequency-selective relay so that the rotor output frequency is between 49.7 and 50.3 cps even though the frequency of the stator input varies from 47.5 to 51.4 cps. An electronic frequency corrector is also described which rectifies the variable-frequency supply and reconverts the dc into ac at the required frequency by means of inverting tubes and a timing unit. Both kinds of equipment are operated by remote automatic control. See also *Engineering* (London), vol. 167, pp. 331–332; April 8, 1949.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 2055

Ultrafax—D. S. Bond and V. J. Duke. (*RCA Rev.*, vol. 10, pp. 99–115; March, 1949.) See also 1203 of May.

621.397.5 2056

Polycast System for TV on U.H.F.—R. M. Wilmotte and P. A. Demars. (*FM-TV*, vol. 8, pp. 26–28, 46; December, 1948.) For the United States uhf television band, the power required to obtain satisfactory coverage at all points within say 30 miles of a single transmitter is prohibitive. "Polycasting" is a suggested alternative, using ten to fifteen 200-watt transmitters within the area, each covering an area of 10 miles radius.

621.397.5:535.88:791 2057

Theater Television—(*Jour. Soc. Mot. Pic. Eng.*, vol. 52, pp. 243–267; March, 1949. Bibliography, pp. 268–272.) Report, in language as little technical as possible, of the Theater Television Committee of the Society of Motion Picture Engineers on the present state of the art.

621.397.5:621.3.09 2058

Attenuation and Phase Distortion and Their Effect on Television Signals—Fuchs and Baranov.(*See* 1871.)

621.397.5:621.315.212 2059

The London-Birmingham Television Cable: Part 2—Cable Design, Construction and Test Results—Stanesby and Weston. (*See* 1857.)

621.397.5:621.396.1 2060

Allocation of Frequencies for the Television Service—McL. (*See* 2047.)

621.397.6:621.385.832 2061

The Graphechon—A Picture Storage Tube—L. Pensak. (*RCA Rev.*, vol. 10, pp. 59–73; March, 1949.) 1949 IRE National Convention paper. Describes the combination, within one envelope, of a cathode-ray tube and an iconoscope, enabling a picture written once on the common screen to be scanned continuously for 1 to 2 minutes. The picture is recorded on the screen by use of the fact that the thin insulating layer on the iconoscope target becomes conductive at the point of impact of a high-velocity electron beam. The picture is read by the conventional iconoscope method. The graphechon was designed originally in connection with the teleran navigation system.

621.397.6:621.395.667 2062

Phase and Amplitude Equalizer for Television Use—Goodale and Kennedy. (*See* 1893.)

621.397.61:621.3.015.3 2063

Standardization of the Transient Response of Television Transmitters—R. D. Kell and G. L. Fredendall. (*RCA Rev.*, vol. 10, pp. 17–34; March, 1949.) It is suggested that picture-monitoring receivers for vestigial-sideband television transmitters should now be standardized on the basis of response to a double-sideband signal modulated by a square wave. Tolerances for vestigial-sideband transmitter performance could then be defined by the shape of the monitor receiver response when the transmitter is modulated by a square wave. Phase-correction networks for reducing the distortion introduced by the vestigial-sideband system and by the restricted bandwidth used are also discussed.

621.397.61-182.3 2064

The WMAL-TV Mobil TV Unit—F. W. Harvey and E. D. Hilburn. (*Communications*, vol. 29, pp. 8–11, 31; March, 1949.) An illustrated description.

621.397.62

Experimental Television Receiver—(? (Franç.), no. 45, pp. 13–16, 21; March, 1 Circuit diagrams and complete details of ceiver using a small electrostatic cathod tube with a screen only 7 cm in diameter capable of receiving either the present 45(or the proposed 819-line transmissions.

621.397.62:621.396.662

Simplified TV Receiver Channel Swit Mechanism—J. A. Hansen. (*Tele-Tech*, v pp. 36–38, 72; December, 1948.) A compac channel superheterodyne tuner which u sliding carriage to carry the tuning elem The sound if is about 21.75 Mc and the v if 26.25 Mc. Performance is discussed.

621.397.743(73)

Television in 1949—Stations and Netw [in the U.S.A.]—(*Tele-Tech*, vol. 7, Decen 1948. Supplement.) Map and brief tabul data of stations now existing or under struction.

621.397.823:629.135

Television Interference by Aircraft—I Cooper. (*Wireless World*, vol. 55, pp. 142- April, 1949.) The use of directive rece antennas reduces the region within whic aircraft causes interference, but a greate round reduction of interference is usuall tained by raising the receiving antenna. moval of the dc component from the rec signal eliminates fluctuations in picture br ness; the changes in contrast due to the i ference are not then so noticeable. The d of a filter to attenuate frequencies betwe cps and 25 cps, to mitigate some disadvant of removing the dc component, is discusse

TRANSMISSION

621.396.61+621.396.621]:551.510.535

Ionosphere Sounding [equipment] Maguer. (*Radio Franç.*, pp. 7–12; April, 1 Description, with complete circuit diagram (a) a transmitter giving 50-μsecond pulses peak power of 1.6 kw and recurrence frequ of 50 per second, and (b) a wide-band su heterodyne receiver with good sensitivity which a cro sweep circuit of simple design g a height-scale accuracy approximating to obtained with a crystal oscillator and its s what complex frequency-division circuits.

621.396.61:621.316.726

Frequency Control in Transmitters—H R. Boosman and E. H. Hugenholtz. (C *munication News*, vol. 9, pp. 21–32; Septem 1947.) Discussion of various methods of taining high stability of frequency and curacy of tuning, including methods invol the principles of frequency-adjustment co and decade tuning. A number of discrimin circuits and circuits suitable for synchron tion with a high harmonic of a control quency are also described.

621.396.61:621.316.726

Telesynchronization with Standard quency—L. Rohde and R. Leonhardt. (F *meldetech. Z.*, vol. 2, pp. 85–90; March, 19 Detailed discussion of methods of phase frequency synchronization, with partic reference to the control of common-w broadcasting transmitters. Methods of deri a control voltage from the phase or freque difference between the standard and the l generator are described. The case where ratio of the standard and local frequencies that of two reasonably small integers is considered.

VACUUM TUBES AND THERMIONIC

621.383

Lead Selenide Photoconductive Cells J. Milner and B. N. Watts. (*Nature* (Londo

163, p. 322; February 26, 1949.) Cells pre-
l by an experimental chemical deposition
ss are found to have a broad sensitivity
mum at 3 to 4 μ. Appreciable sensitivity
so far only been exhibited at low tem-
ures.

83 **2073**
ismuth Sulphide Photocells—B. T. Kolo-
s. (*Zh. Tekh. Fiz.*, vol. 19, pp. 126–131;
ary, 1949. In Russian.) The photo cells
nade with synthetic Bi_2S_3. Various ex-
nental characteristics are plotted. This
of cell is particularly suitable for auto-
c control applications.

83 **2074**
hotoelectric Multipliers—S. Rodda. (*Jour.
Instr.*, vol. 26, pp. 65–70; March, 1949.)
ission of fundamental principles, emissive
rials, electrode shapes, current fluctua-
dark current, fatigue phenomena, prac-
circuits, applications, etc.

83:535.215 **2075**
**n the Interpretation of Observations on
Photoelectric Voltages with Intermittent**
—Gorter, Broer, and Snoek. (*See* 1916.)

83:535.215 **2076**
**n Photoelectric Voltages in Light-Absorb-
Materials**—Snoek and Gorter. (*See* 1915.)

83:621.396.645 **2077**
**he Development of a Photoelectric A.C.
ifier with A.C. Galvanometer**—Milatz and
nbergen. (*See* 1908.)

85 **2078**
tandard Valves [Book Review]—Standard
hones and Cables, 1947, 328 pp., 15s.
ire (London), vol. 163, p. 387; March 12,
) Details and technical data of ordinary
, cathode-ray tubes, cold-cathode relays,
tubes and disk-seal uhf triodes manufac-
by the above firm. Brimar tubes are not
led.

85 **2079**
**ew Miniature American-Type Valves
in France**—M. Leroux. (*Radio Prof.*
), vol. 18, pp. 6–9 and 18–20, 23; March
April, 1949.) Full technical details, with
tional characteristics and practical cir-
The tubes include pentodes 6AG5, 6AU5,
, 6AK5, output tubes 6AQ5 and 6AK6,
e-diode triode 6AT6, diode 6AL5, double
6J6, high-μ triode 6J4, rectifier 6X4, and
tubes 12BE6, 12BA6, 12AT6, and
6 which are identical with the correspond-
volt tubes except that their heater cur-
0.15 ampere at 12.6 volts.

85 **2080**
New Series of Small Radio Valves—G.
and F. Prakke. (*Philips Tech. Rev.*, vol.
289–295; October, 1946.) Discussion of
A" series or "Rimlock" tubes, 22 mm in
ter, for which overheating of the cathode
ided by joining the glass or metal bulb
flat glass base with a glaze or cement
becomes plastic at a comparatively low
rature.

85 **2081**
ew Post-War German Valves—(*Radio*
(Vienna), vol. 25, p. 249; April, 1949.)
particulars of 15 tubes, including type of
nd of body and, in some cases, heater
e and current. Output data are included
new developments: UEL71 and VEL11,
tetrode/pentodes, and UL2, a 1.5-watt
t pentode.

85 **2082**
**ries of Modern Valves for F.M. Broad-
g and for Television**—J. Becquemont.
Élec., vol. 29, pp. 145–151; April, 1949.)
ssion of methods of achieving small inter-electrode distances, high electron densities, high anode dissipation, low reaction capacitance and output impedance, high gain and complete separation of input from output. The use of Dilver-P, an alloy comparable to kovar, together with a glass of the same expansion coefficient has made possible the construction of a series of tubes in which the various electrodes are carried on concentric cylinders, which form the base connections. The operation of welding the Dilver-P cylinders to the intermediate glass rings to form the tube base is carried out in a high-temperature furnace, using a graphite mandrel. This construction gives short connections to the electrodes. Illustrations are given of a 500-watt triode and a 2-kw tetrode with anode cooling fins.

621.385 **2083**
Theory and Applications of Trochotrons—H. Alfvén; L. Lindberg; K. G. Malmfors; T. Wallmark; and E. Åström. (*Kungl. Tekn. Högsk. Handl.*, (Stockholm), no. 22, 106 pp; 1948. In English.) Includes 5 separate but coordinated papers. The first, "On Trochoidal Electronic Beams and Their Use in Electronic Tubes (Trochotrons)," by Alfvén, is an integrating paper. The equations of motion of charged particles in a nearly homogeneous magnetic field perpendicular to a nearly homogeneous electric field are discussed, and the properties of the resulting trochoidal beams are considered in detail. The arrangement of a simple trochotron is described. The cathode is an electron gun, placed between an L-shaped anode and a straight "rail" electrode at nearly cathode potential. The electron beam travels between the "rail" and a number of "boxes." The anode forms one side of the last box, while the side of the first box remote from the cathode is connected to the rail. Electrodes forming sides of the boxes, at right angles to the rail, are called spades; those forming sides parallel to the rail are called plates. All spades and plates, and the anode, are normally at a potential of +200 volts relative to the cathode, except the spade connected to the rail. The beam may enter the first box and be collected by the first plate. The beam can be made to enter one of the other boxes by altering the voltage of the corresponding spade. For earlier work, see 3800 of 1945 and 1205 of 1948.

The second paper, "Design and Properties of Trochotrons," by Wallmark, discusses the trochotron in greater detail, with illustrations of the way in which the beam behaves when some electrode voltages are changed by steps. Operating conditions, the size of the electrode system, electrode materials, etc., are considered.

The third paper, "Design of Trochotron Circuits," by Lindberg, discusses the way in which external pulses may be used to operate each spade as a self-locking switch. Applications to counters, chronoscopes, and pulse-time modulation are also discussed.

The fourth paper, "Experimental Investigation on an Electron Gas in a Magnetic Field," by Åström, discusses experimental results concerning currents which flow to electrodes whose potential is negative relative to the cathode. These "negative currents" are always accompanied by noise. They occur even if there is no alternating voltage on the electrodes. This phenomenon is related to cut-off effects in magnetrons.

The fifth paper, "On the Instability of an Electron Gas in a Magnetic Field," gives theory primarily proposed as an explanation of Åström's results, but it may also throw light on the origin of solar and cosmic noise.

621.385 **2084**
Valves with Resistive Loads—S. W. Amos. (*Wireless Eng.*, vol. 26, pp. 119–123; April, 1949.) Expressions are derived for the maximum undistorted output that can be delivered to a purely resistive load by a tube with given high-voltage supply. Although the expressions are derived for the cathode-follower circuit, many of the formulas may be applied to tubes with the load in the anode circuit.

621.385 **2085**
Some Recent Developments in the Technique of Radio Valve Manufacture—J. W. Davies, H. W. B. Gardiner, and W. H. Gomm. (*Proc. Inst. Mech. Eng.* (London), vol. 158, no. 3, pp. 352–363; 1948. Discussion, pp. 364–368.) A general account of the mechanical aspects of the manufacture of large transmitting and high-frequency tubes, with particular attention to the making of spiral, squirrel-cage, and planar grids, glass-to-metal joints of the annular type, using chucks for holding the component parts, and the assembly and alignment processes.

621.385:061.3 **2086**
Digests of Papers Presented at Conference on Electron Tubes—(*Elec. Eng.*, vol. 67, pp. 589–600; June, 1948.) Authors' summaries of most of the papers read are given. Full texts are being published in the *Proceedings of the Conference on Electron Tubes* (price $3). The *Report on Electron Tube Survey* (price $2) prepared by an AIEE Committee, contains data on which much information discussed at the conference was based. Both are obtainable from AIEE Order Department, 33 West 39 Street, New York 18, N. Y.

621.385:621.396.822 **2087**
The Noise Factor of Grounded-Grid Valves—A. van der Ziel and A. Versnel. (*Philips Res. Rep.*, vol. 3, pp. 255–270; August, 1948.) A mathematical treatment of grounded-grid tubes in which part of the output noise current also flows in the input circuit, giving partial noise suppression. Triodes, pentodes, and secondary-emission tubes are considered, and the effects of circuit and dielectric losses, transit time, field inhomogeneities, partition noise, and secondary-emission noise are discussed.

Measurements made on tubes for $\lambda = 7.25$ meters agree with theory for curves of noise factor against antenna resistance, and for noise resonance curves. Loose antenna coupling favors a low noise factor, but complete noise suppression is impossible. Suppression is most effective for grounded-grid triodes over a narrow frequency band, so that this type of tube is less useful for wide-band working. The noise factor for secondary-emission tubes is much greater than that for triodes. See also 249 of February.

621.385.029.63/.64 **2088**
Some Slow-Wave Structures for Traveling-Wave Tubes—L. M. Field. (Proc. I.R.E., vol. 37, pp. 34–40; January, 1949.) The gain per unit length of a traveling-wave tube is determined by a structure factor. Four types of structure are compared, namely (a) helix, (b) disk-loaded rod, (c) apertured-disk, and (d) helical waveguide. Amplifier performance of helix tubes at 10,000 Mc and wide-tuned oscillator performance (1.5 to 1) with second-harmonic output around 20,000 Mc are considered. Design, construction, and performance of a 10,000-Mc tube of type (b) are also discussed.

621.385.029.63/.64 **2089**
Effect of the Transverse Electric Vector in the Delay Line of the Traveling-Wave Valve: Part 1—O. Doehler and W. Kleen. (*Ann. Radioélec.*, vol. 4, pp. 76–84; January, 1949.) The hypotheses used in different theories of the linear behavior of the traveling-wave tube are briefly reviewed and the effect of the transverse electric field of the delay line on the interaction between the beam and the wave is examined. This effect appears to have been neglected hitherto. The radial field causes five waves to be excited instead of the three found on the

hypothesis of a purely longitudinal field; the gain of the amplified waves can be appreciably increased by the action of the radial field. Particular types of delay line where such effects predominate are discussed and the change of gain due to the space charge and to the absorption of electrons by the delay line is studied. To be continued. See also 1543 of June (Pierce).

621.385.032.216 2090

Fluctuation Effects of Emission from Oxide-Coated Surfaces—R. P. Bien and Yang Yo-Han. (*Science Rec.* (Nanking), vol. 2, pp. 65–70; October, 1947. In English.) Fluctuations observed bear no relation to heating current but can be explained statistically.

621.385.032.216 2091

Thermionic Emission from Oxide Coated Cathodes—D. A. Wright. (*Proc. Phys. Soc.*, vol. 62, pp. 188–203; March 1, 1949.) Discussion of an experimental investigation of the emission from Ba/Sr oxide cathodes and from thoria cathodes under pulsed and dc conditions. Semiconductor theory can explain the emission and conductivity of cathode coatings. The part played by the interface layers between the coating and the metal to which it is applied is also considered. For earlier work see 606, 1980, and 1981 of 1948.

621.385.032.216 2092

Poisoning in High-Vacuum Oxide-Cathode Valves—G. H. Metson and M. F. Holmes. (*Nature* (London), vol. 163, pp. 61–62; January 8, 1949.) The ionizing action of the current after leaving the cathode surface is suggested as the prime factor leading to cathode deterioration. Results of life tests on two batches of a particular type of pentode are discussed. Those arranged as triodes with anodes strapped to screen and suppressor grids and primed with 200 volts failed after 500 to 1,000 hours, whereas those connected as diodes, with control grids acting as collectors and +2 volts applied between grid and cathode to give a cathode current of about 12 milliamps showed little deterioration after 6,000 hours.

621.385.032.216 2093

Resistance of Oxide Cathode Coatings for High Values of Pulsed Emission—W. E. Danforth and D. L. Goldwater. (*Jour. Appl. Phys.*, vol. 20, pp. 163–173; February, 1949.) The potential variation of fine ribbon probes embedded in standard BaO or SrO coatings was observed for 19 tubes. Potential gradients were found adequate to admit dielectric breakdown as a cause of sparking. The resistance of SrO cathodes is several times that of BaO or mixed-oxide cathodes. Superposition of dc upon pulsed emission causes a marked decrease in resistance.

621.385.032.216 2094

Work Functions and Conductivity of Oxide-Coated Cathodes—G. W. Mahlman. (*Jour. Appl. Phys.*, vol. 20, pp. 197–202; February, 1949.)

621.385.032.42 2095

A New Air-Cooling System for Transmitting Valves—W. L. Vervest. (*Communication News*, vol. 9, pp. 92–96; April, 1948.) See also 2673 of 1948 (de Brey and Rinia).

621.385.032.5:666.1.037.5 2096

The Electrode Leads of Transmitting Valves—E. G. Dorgelo. (*Communication News*, vol. 9, pp. 38–44; December, 1947.) Survey and discussion of various methods of sealing.

621.385.2:621.396.822 2097

Noise Spectrum of Temperature-Limited Diodes—D. B. Fraser. (*Wireless Eng.*, vol. 26, pp. 129–132; April, 1949.) A simple and completely general derivation of the formula for the fluctuation currents in plane and cylindrical diodes, with explicit allowance for the transit time. When transit time is negligible, the formula reduces to that of Schottky.

621.385.2:621.396.822 2098

Noise Spectrum of a Diode with a Retarding Field—J. J. Freeman. (*Jour. Res. Nat. Bur. Stand.*, vol. 42, pp. 75–88; January, 1949.) A general expression is derived for the spectrum generated by the random emission of electrons having arbitrary trajectories within a waveguide. A numerical solution is obtained for the potential distribution within a plane diode; results are shown as a series of curves, and compared with von Laue's results. The equivalent mean-square fluctuation current due to the space charge within a diode is deduced for (a) linear potential distribution, and (b) the distribution which occurs near the beginning of the retarding field. In case (a), the equivalent noise temperature of the diode conductance is equal to the cathode temperature.

621.385.2.032.216 2099

Some Characteristics of Diodes with Oxide-Coated Cathodes—W. R. Ferris. (*RCA Rev.*, vol. 10, pp. 134–149; March, 1949.) The Epstein-Fry-Langmuir equation for the space-charge current in a plane diode and the Boltzmann equation for the retarding-field current are used to obtain a set of universal characteristic curves for plane diodes; $V.|e/kT|$ is taken as the abscissa. These curves agree with experimental measurements when a series resistance is assigned to the oxide cathode. Universal curves of incremental conductance and tables of functions used are given.

621.385.3.029.64 2100

Electronics of Ultra-High-Frequency Triodes—R. R. Law. (Proc. I.R.E., vol. 37, pp. 273–274; March, 1949.) An empirical relation is deduced for anode efficiency as a function of frequency, voltage, and interelectrode spacing.

621.385.3.032.29:621.317.335.2† 2101

Triode Interelectrode Capacitances—E. E. Zepler and J. Hekner. (*Wireless Eng.*, vol. 26, pp. 53–58; February, 1949.) The variations of grid-to-cathode and grid-to-anode capacitance with working conditions were investigated experimentally. The effects of the mutual conductance, amplification factor, and supply voltages are shown graphically. A theory is given which is in fair agreement with these results.

621.385.032.29:621.317.335.2† 2102

Interelectrode Capacitance of Valves—B. L. Humphreys and E. G. James. (*Wireless Eng.*, vol. 26, pp. 26–30; January, 1949.) Discussion of measurements made under different operating conditions, on two types of tube—DET 22 and E 1714—whose active elements are all cylindrical and coaxial. The measurements were made on a rf bridge at a frequency of 1 Mc. The increase of grid-to-cathode capacitance with increasing anode current was much greater than that expected theoretically and depended greatly on the grid-to-cathode geometry. Anode-to-grid capacitance decreased very slightly with increasing anode current.

621.385.832:621.397.6 2103

The Graphechon—A Picture Storage Tube—Pensak. (*See* 2061.)

621.396.615.142:621.316.726 2104

Frequency Stabilization of V.M. Valves—H. Borg. (*Wireless Eng.*, vol. 26, pp. 59–73; February, 1949.) A simplified discussion of general principles mainly in connection with continuous-wave microwave oscillators. Frequency-control systems are considered in which an error voltage, generated between a standard reference frequency source and the oscillator to be stabilized, is used to correct frequency variations of the oscillator. An application described in which the frequency of a v tube at 9,360 Mc is stabilized by compari with a crystal, giving short-term stabil relative to the crystal, of the order of + cps.

621.396.615.142.2 2

Valves for Communication on Frequen above 1,000 Mc/s: Part 1—H. Schnit (*Fernmeldetech. Z.*, vol. 2, pp. 51–56; Febru 1949.) A general outline of the principles klystrons of the normal and the reflex ty with a short account and illustration of th chamber klystron of the German Post Off this gives an output of 100 watts on a wa length of 9 cm.

621.396.645:537.311.33:621.315.59 2

Germanium—Important New Semic ductor—Dunlap, Jr. (*See* 1968.)

621.396.645:537.311.33:621.315.59 2

Some Novel Circuits for the Three-T minal Semiconductor Amplifier—Webs Eberhard, and Barton. (*See* 1905.)

621.396.645:537.311.33:621.315.59 2

The Double-Surface Transistor—J. Shive. (*Phys. Rev.*, vol. 75, pp. 689–6 February 15, 1949.) Emitter and collec point-contacts bear on opposite faces of a t wedge of Ge and a third contact of lar area is provided on the base of the wed Separation between the points should exceed 0.1 mm. Families of curves are gi which facilitate correct choice of the operating voltage and give complete inforr tion for determining the dynamic input a output impedances and the forward and ba ward transfer impedances about any selec operating point. An explanation of the act of this type of transistor is given. See also of April (White) and back references and 2 below.

621.396.645:537.311.33:621.315.59 2

Investigation of Hole Injection in Trans tor Action—J. R. Haynes and W. Shockl (*Phys. Rev.*, vol. 75, p. 691; February 15, 194 The impedance changes at the collector po of a transistor were investigated by applyi an intermittent potential to multiple emit points and observing the resulting probe-c rent variations. The results are explained terms of the movement of positive partic with a mobility of about 1.2×10^3 cm/sec v/cm.

621.383 2

Photoelectric Cells in Industry [Book R view]—R. C. Walker. Pitman and Sons, Lo don, 501 pp., 40s. (*Electronic Eng.* (Londo vol. 21, p. 68; February, 1949.) Typical uses such cells are discussed from the point of vi of the practical man whose purpose is to u electrons rather than to theorize about them.

MISCELLANEOUS

621.39 2

What the S.C.E.L. [Signal Corps Enginee ing Laboratories] is Doing—H. A. Zal (*FM-TV*, vol. 9, pp. 13–22; February, 194 A general survey of work in many fields. D velopment work is the primary task, but bas and applied research work related to the mi tary effort is also undertaken.

621.396 2

Radio Progress during 1948—(Proc. I.R. vol. 37, pp. 286–322; March, 1949.) A gene review with a bibliography of 855 reference

621.396 Popov 2

Alexander S. Popov—J. B. Thornto (*Wireless Eng.*, vol. 26, pp. 141–142; Apr 1949.) Comment on 1842 of 1948 (G. W. O. H.